Remarks

Claims 1-24 are pending. Claims 1, 6-10 and 13-22 stand rejected as unpatentable over Hollenback in view of Talambrias and further in view of Dayton. Claims 2, 3, 11, 12, 23 and 24 stand rejected as unpatentable over Hollenback in view of Talambrias and further in view of Dayton and further in view of Smith. Claims 4 and 5 stand objected to as being dependent upon a rejected base claim. Applicants traverse the rejections.

In maintaining rejection of the claims, the Examiner asserts with respect to claim 1 that "transconductance amplifiers are conceptual models while operational amplifiers are physical devices." The Examiner similarly asserts with regard to claim 8 that a voltage amplifier is easily modeled as a transconductance amplifier.

Applicants enclose a series of references that illustrate that transconductance amplifiers are, in fact, physical devices that are utilized in circuit design. The references are:

- "Transconductance Amplifiers Simplify Wideband Techniques," Application Note 692, Maxim Integrated Products, March 15, 2000 confirms the existence of transconductance amplifiers as physical devices, discusses their application, and notes that "limited performance in transconductance amplifiers has hampered their acceptance for years, with exception of the few applications tailored to their capabilities. See page 1 of Application Note 692.

- "Application of the Operational Transconductance Amplifier (OTA) to Voltage-controlled Amplifiers and Active Filters," W. Grise, Department of IET Morehead State University, http://et.nmsu.edu/~etti/winter98/electronics/grise/wrg.html also confirms the existence of "presently available commercial OTA's." See page 1 of Grise. This paper also notes that the OTA is "a voltage-controlled current source (VCCS), which is in contrast to the conventional op-amp, which is a voltage-controlled voltage source (VCVS)." See page 2 of Grise. Also note National Semiconductor's MB13600 is discussed as a commercially available OTA. See page 2 of Grise.

- "LM13700 Dual Operational Transconductance Amplifiers with Linearizing Diodes and Buffers;" National Semiconductor, June 2004, is one version of a commercially available OTA.

These references establish that transconductance amplifiers are physical devices that are commercially available. In contrast to the Examiner's assertion that transconductance

amplifiers are merely conceptual models, these references show that they are physical devices that are distinct from operational amplifiers, as argued previously by Applicants. In view of these remarks, the Examiner is requested to withdraw the final rejection of the claims.

The application is considered in good and proper form for allowance, and the Examiner is respectfully requested to pass this application to issue. If, in the opinion of the Examiner, a telephone conference would expedite the prosecution of the subject application, the Examiner is invited to call the undersigned attorney.

Respectfully submitted,

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Date: May 22, 2006

Transconductance Amplifiers Simplify Wideband Techniques

This article describes the unique architecture used in the MAX435/MAX436 transconductance amplifier and how this applies to traditional applications as well as new ones. Sample circuits are shown using the MAX435/MAX436 as a phase splitter, an impedance transformer, a coaxial cable driver, and as a twisted pair cable driver for distances over 5000 feet.

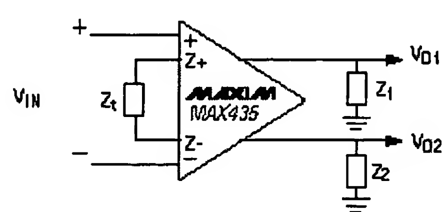
Limited performance in transconductance amplifiers has hampered their acceptance for years, with exception of the few applications tailored to their capabilities. But two new products from Maxim promise to widen the scope of such amplifiers. The Maxim parts offer better specs for established circuits, and their unique architectures offer the prospect of entirely new applications.

MAX435/MAX436 amplifiers are open-loop devices that provide accurate gain without feedback. VOUT/VIN gain is the product of an internal current gain (4 ±2.5% in the MAX435; 8 ±2.5% in the MAX436), and the ratio of an output impedance ZL to the user-connected "transconductance network" Zt (**Figure 1**). Zt is a 2-terminal network connected across the amplifier's Z+ and Z- terminals. The MAX435 has differential outputs, and the MAX436 has a single-ended output.

TWO EQUATIONS:

$$V_{O1} = K^* \left(\frac{Z_1}{Z_t}\right) V_{IN}$$

$$V_{O2} = -K^* \left(\frac{Z_2}{Z_t}\right) V_{IN}$$



*K = ± 2.5% (MAX435), 8±2.5% (MAX436)

GAIN IS SET BY A RATIO OF TWO IMPEDANCES AND AN INTERNAL CURRENT GAIN FACTOR (K).

Figure 1.*Simple equations and freedom from instability ease the application of transconductance amplifiers.*

Because ZL or Zt (or both) can be frequency-shaping networks, the ZL/Zt ratio can implement some interesting transfer functions. A resistor ratio (times the internal current gain) simply sets a desired voltage gain. Replacing ZL with a parallel-RC network produces a lowpass response, and replacing Zt with a series-RC network produces a highpass response. Combining the

parallel-RC ZL and series-RC Zt produces a bandpass filter. Or, by replacing Zt with a crystal or series-LC network you can create a high-Q tuned amplifier.

Each of these configurations is elevated to new levels of performance by the amplifiers' high speed: the MAX435 has a 275MHz bandwidth with 800V/µs slew rate, and the MAX436 has a 200MHz bandwidth with 850V/µs slew rate. Both offer 18ns settling times (±1%) for 0.5V step inputs, and both feature exceptional CMRRs of 53dB at 10MHz. Both have fully differential, symmetrical, high-impedance inputs. Input offset voltages (300µV typical) are much lower than those of most high-speed op amps.

The secret of high speed lies in the MAX435/MAX436 architecture. Consider the MAX435 (**Figure 2**). With zero volts across VIN+ and VIN-, the currents from I1 and I2 are mirrored and multiplied, producing 12mA in Q1 and Q2. These currents each match 12mA from a current source in the output stage, producing a zero differential output at IOUT+ and IOUT-.

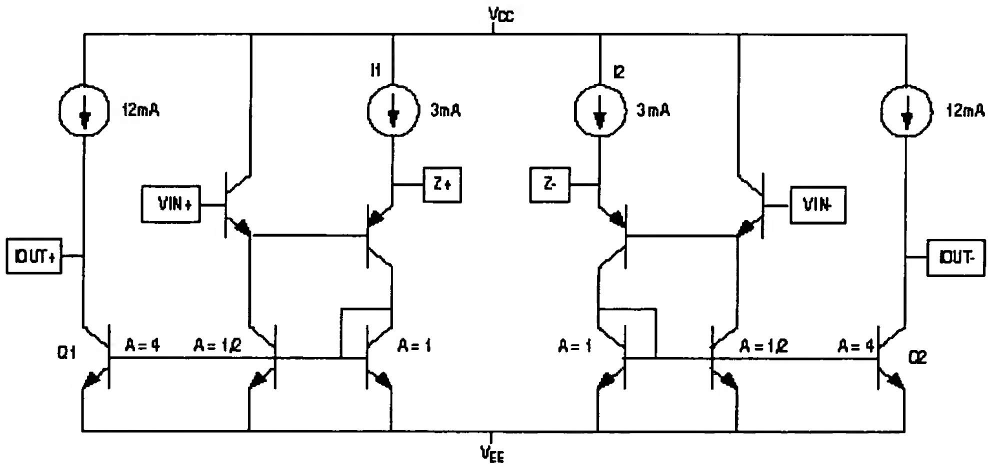


Figure 2. *This simplified schematic shows basic circuitry in the MAX435 differential-output transconductance amplifier. An external resistor (R_{SET}) controls the four current sources, and its nominal value of 5.9k   produces the current levels shown.*

Connecting a positive differential voltage across VIN+ and VIN- diverts some of the I1/I2 current through Zt (connected between Z+ and Z-), causing an imbalance in the Q1/Q2 currents. The result is a net differential output current at IOUT+ and IOUT-. Time delays are very short because the signals propagate as steered currents (rather than voltages), and because all stages in the signal path receive substantial bias currents. The following applications are made possible by these and other special capabilities in the MAX435/MAX436 amplifiers.

Because MAX435 and MAX436 outputs are high-impedance current sources, you can create a summing amplifier simply by tying two or more outputs together. No additional components are required except a load resistor to develop the output voltage. Another intrinsic function is that of phase splitter-the MAX435 differential outputs provide inverted and non-inverted (0° and 180°) versions of the input signal.

As phase splitter, the MAX435 offers a convenient, single-IC differential drive for balanced transmission lines (**Figure 3**). The IC's excellent common-mode rejection (90dB at dc; -53dB at 10MHz) assures reliable transmissions.

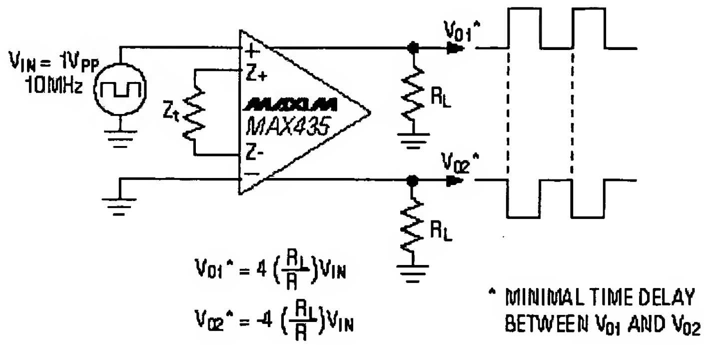
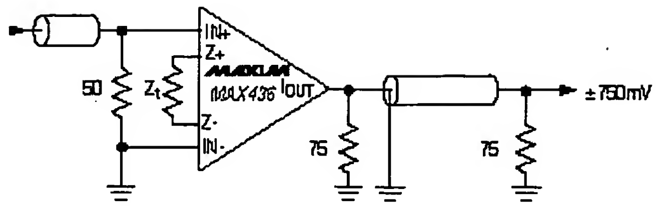
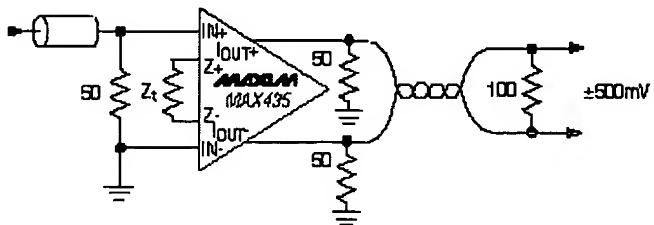


***Figure 3.** Differential outputs make the MAX435 a convenient single-package phase splitter.*

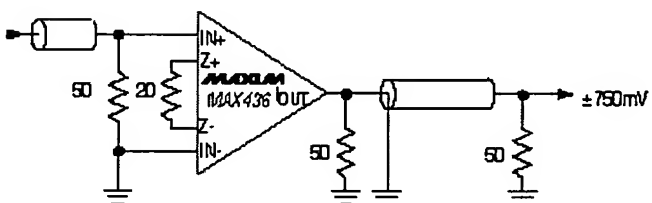
The amplifiers' high-impedance inputs and outputs allow them to operate as monolithic impedance transformers (**Figure 4**). The high-impedance, true-differential inputs (800kW typical) let you connect any reasonable value of input termination resistance. Similarly, the current-source outputs have a relatively high source resistance (3.2kW typical) that lets you connect any reasonable value of load resistance.



50Ω → 75Ω



50Ω → BALANCED PAIR



50Ω → 50Ω, GAIN = 20dB

Figure 4.*Independent settings for output current and load resistance enable MAX435/MAX436 amplifiers to act as impedance transformers. Supply voltages are ±5V, and the R_{SET} resistors (between the amplifiers' I_{SET} terminals and ground) are 5.9k .*

The main advantage of these circuits over magnetic transformers is in their low-end frequency response, which extends to dc. Baseband video, for example, has frequency components ranging from 4.5MHz to below 60Hz. A line transformer with flat frequency response over that range would be very bulky and expensive! Flexibility is another advantage for the IC approach; by changing one or two resistors you can match the transmitter and receiver to a variety of cables in the same system.

As another illustration of the need for impedance matching, coaxial cables for high-speed signals must be carefully terminated in their characteristic impedance to ensure maximum power transfer and minimum distortion. To obtain optimum performance from 50W cable, therefore, you must terminate each end of the cable with 50W.

Further description
Voltage-mode amplifiers have low output impedance, so they require a series-resistor interface to coaxial cable. But MAX435/ MAX436 amplifiers have high-resistance current-source outputs that require a parallel connection of the termination resistor (i.e., in shunt with the cable). Note that back-terminating the cable this way reduces the circuit voltage gain by half **(Figure 5).**

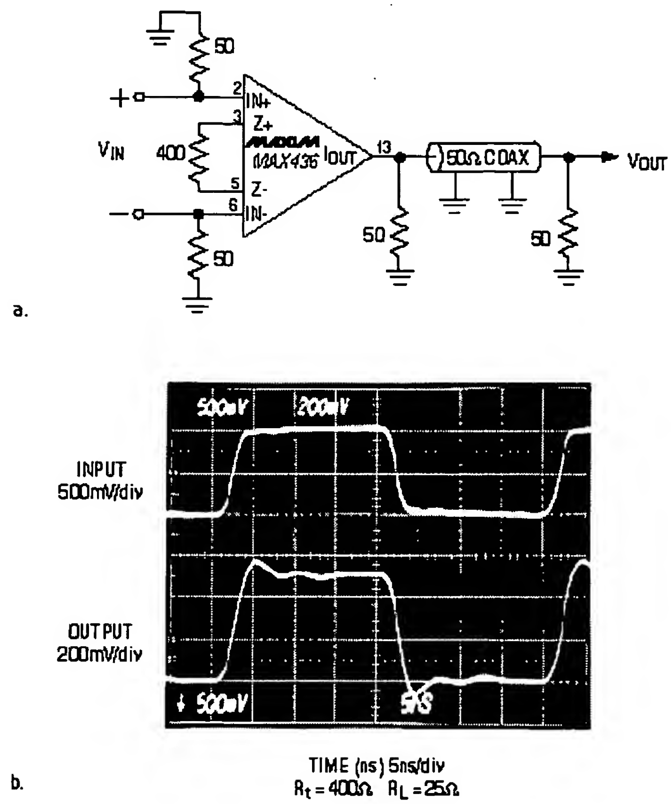


Figure 5. As a coaxial-cable driver (a), the MAX436 transconductance amplifier handles fast pulses with minimal overshoot and ringing (b).

MAX435/MAX436 amplifiers offer the user several "control handles." For top performance in this application and others, you should be aware of the amplifiers' shutdown capability, their adjustable load-current limits, and the factors that affect their dc accuracy.

First, the internal current sources are controlled by an external resistor (RSET) connected between the ISET terminal and the V- supply voltage (Figure 2). Both amplifiers operate on ±5V. The standard RSET value for which all specifications are guaranteed is 5.9kW, and this value sets the limit for maximum IOUT: ±20mA for the MAX436, and ±10mA per output for the MAX435. By connecting a larger-valued RSET, you can reduce the amplifiers' supply current and power dissipation (along with the maximum IOUT).

You can also increase the output current by decreasing RSET, but be careful to ensure that the higher current does not combine with a particular operating condition to exceed the package power-dissipation rating. Removing RSET altogether provides a partial shutdown of the amplifier. Without RSET, the room-temperature supply currents (normally 35mA) drop to 450µA ±25% for the MAX435 and 850µA ±25% for the MAX436.

DC accuracy in the MAX435 and MAX436 is affected by the input offset voltage (VOS), the output offset current (IOS), and tolerance on the internal current gain K, as well as tolerance on the external impedances Zt and ZL. VOS is caused by a VBE mismatch at the input stage (like the VOS in bipolar voltage amplifiers), and is measured between the Z+ and Z- terminals-with Zt removed and the inputs (IN+ and IN-) grounded. VOS produces a small error current in Zt during normal operation. Multiplied by K, it produces an output error current, even with no differential input voltage applied.

IOS is a separate and independent output error that is caused by imperfectly matched devices in the output current mirrors. Though measured under the same conditions as the VOS measurement, IOS does not vary with input voltage. Combining the IOS and VOS effects yields a net error in output voltage. The MAX435's differential output error VERR(DIFF), for instance, is the sum of each output error:

VERR(DIFF) = (VERR+) - (VERR-), where

VERR+ = (RL+)[(IOS+) + K(VOS/Rt)], and VERR- = (RL-)
[(IOS-) - K(VOS/Rt)]. IOS is -20µA typical (±100µA max),
and VOS is 0.3mV typical (3.0mV max).

Similarly for the MAX436,

VERR = (RL)[IOS + K(VOS/Rt)],
where IOS is 6µA typical (±100µA max), and VOS is 0.3mA typical (3mA max).

Twisted-pair video

The MAX435 and MAX436 amplifiers provide a differential-out/differential-in combination that is well suited for one-way transmission of video signals over a twisted-pair cable (**Figure 6**). As a bonus, the MAX436 Zt network provides a means for line equalization and gain adjustment.

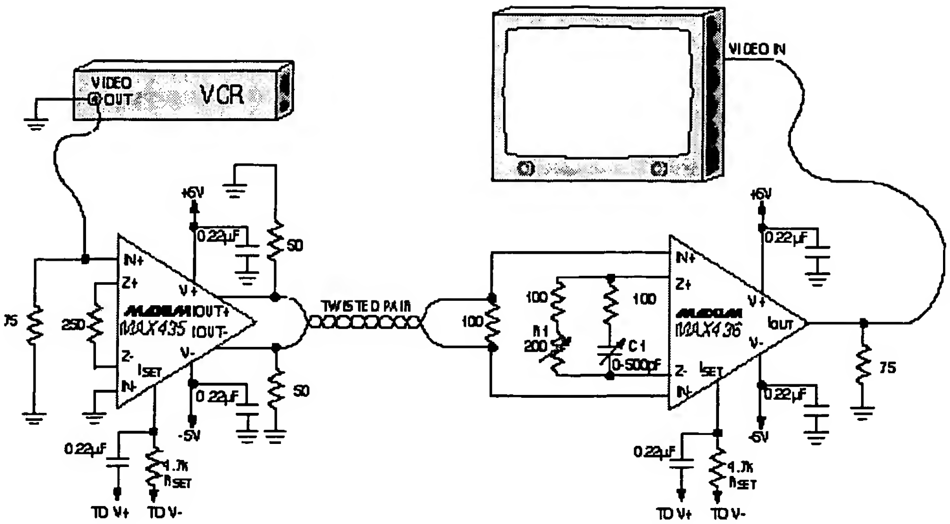


***Figure 6.** Two transconductance amplifiers and a twisted-pair cable transmit baseband video for 5000 feet or more.*

Replacing coaxial cable with twisted-pair cable saves cost in many applications that don't require the higher bandwidth of coax. These applications have initially included LANs and LONs (local area networks and local operational networks). But twisted-pair cable is more compact than coaxial cable, and the miles of unused twisted-pair cabling that already reside in the phone systems of existing buildings may inspire additional applications. Baseband (composite) video can be transmitted over these cables as far as 5000 feet, with surprising quality.

Twisted-pair video transmission works best with a single channel of baseband video. Many applications require such transmissions within a building; an obvious example is the separate video channels routed from individual surveillance cameras back to a security office. Other closed-circuit TV (CCTV) systems are found in retail stores, supermarkets, airports, and schools.

Twisted pairs resist differential noise pickup; because a pair is twisted, any differential current induced by an interfering EM field in one loop gets cancelled in the following loop. Common-mode noise, on the other hand, must be rejected by a balanced (differential) circuit at the receiver. Twisted-pair cables must also be terminated in their characteristic impedance to minimize the reflections caused by line discontinuities.

For twisted pairs exceeding 200 feet (approximately), bandwidth falls short of the typical baseband-video bandwidths (4MHz to 5MHz). But these cables are satisfactory for baseband video if you equalize your receiver, provide an NTSC monitor with automatic gain compensation, and choose quality (wideband) cable.

Stranded and unstranded wires exhibit similar bandwidths, but the highest-bandwidth cables are unshielded, and have insulation of low dielectric constant between the conductors.

Polyethylene or polypropylene insulation is recommended for new installations. For twisted-pair video transmissions under 1000 feet, use common 24AWG telephone wire. For longer distances, you can improve the video fidelity by using larger wire.

The differential-output MAX435 of Figure 6 eliminates the need for a balun (balanced-to-unbalanced) transformer or the two-driver alternative-one single-ended inverting driver and one single-ended non-inverting driver. The MAX435 drives the balanced twisted-pair cable from a ground-referred input signal (in this case, from a VCR's VIDEO OUT baseband signal).

At the driver end of the cable, each conductor is terminated with a 50W resistor to ground. The resulting 100W between conductors is an appropriate match for the cable's characteristic impedance. A mismatch can degrade the video, but it cannot affect amplifier stability because the MAX435 has no feedback. Output amplifiers are ±0.5V.

At the receiver end, a MAX436 amplifier converts the balanced input channel to a single-ended output. Again, the proper line termination is 100W between cable conductors at the IN+, IN- inputs. The Zt impedance network across Z+ and Z- adds adjustable gain (approximately 6dB) to compensate for a 6dB loss introduced by the termination resistors. The network's adjustable capacitor also provides line equalization (frequency compensation) if required. Load resistance is 50W, consisting of the 75W resistor in parallel with 150W at the monitor's input port.

Test results

Operating with 500 feet of inexpensive, 22-gauge, twisted-pair burglar-alarm cable (approximately 4¢ per foot), the Figure 6 circuit attenuates the baseband video's 3.58MHz colorburst frequency about 6dB (**Figure** 7). Despite the distortion, no degradation of color saturation was observed at the NTSC monitor used in this test. No degradation was expected, however; this monitor compensates for signal attenuation by calibrating automatically against test patterns in the vertical interval test signal (VITS).

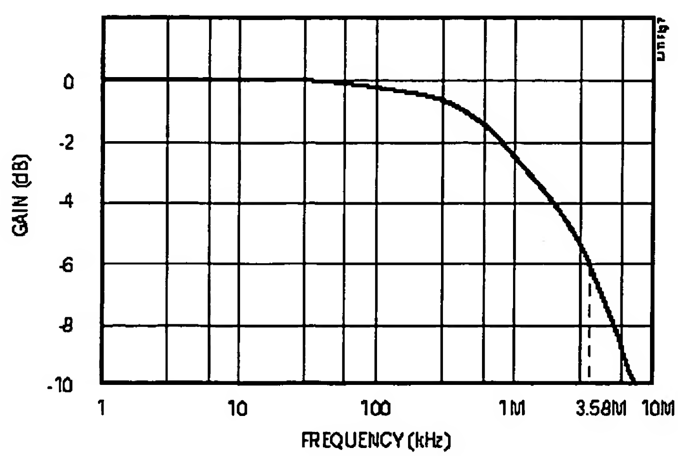
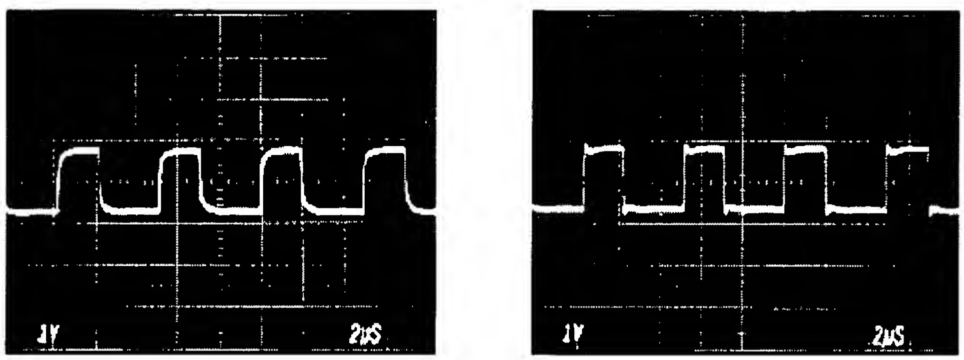


Figure 7. *Inexpensive burglar-alarm cable (twisted pair, 500 feet, 22AWG) attenuates the 3.58MHz colorburst frequency of baseband video by 6dB.*

The monitor's automatic loss equalization is robust; it compensates for colorburst attenuation as high as 10dB, displaying an excellent picture with no noticeable color fading or loss of horizontal resolution. Further attenuation, however, produces poor chroma and a horizontal fuzziness that makes it difficult to read displayed text.

Under that condition you can still achieve compensation via adjustments at the MAX436 Zt network: R1 adjusts brightness by boosting the overall gain to compensate for ohmic losses, and C1 introduces a pole/zero pair in the receiver circuit, which adjusts for color by extending the channel bandwidth. Because compensation is introduced at the receiver, you can simply view the display and adjust for the best picture. Before-and-after waveforms show the result of this equalization (**Figure 8**).



8a. BEFORE EQUALIZATION   8b. AFTER EQUALIZATION

Figure 8. *These before-and-after waveforms show the effect of adjusting for optimum brightness and color via R1 and C1 (Figure 6), while observing the monitor display.*

Next, consider the Figure 6 circuit operating with 1000 feet of twisted-pair telephone cable. The test setup included a length of unused twisted pair in a trunk cable between two Maxim buildings, two jumper connections in the phone-patch room, and additional twisted-pair cable that was routed through hallways to complete the transmission path.

This system easily transmitted baseband video from a VCR, producing an excellent picture with R1 and C1 at their nominal settings (no equalization required). High noise immunity was illustrated by coupling 60Hz common-mode noise to the line (**Figure 9**). The MAX436 CMRR (60dB at 60Hz) removed this noise with no evidence of beating in the display. On the other hand, driving the cable in an unbalanced mode produced poor results as expected.

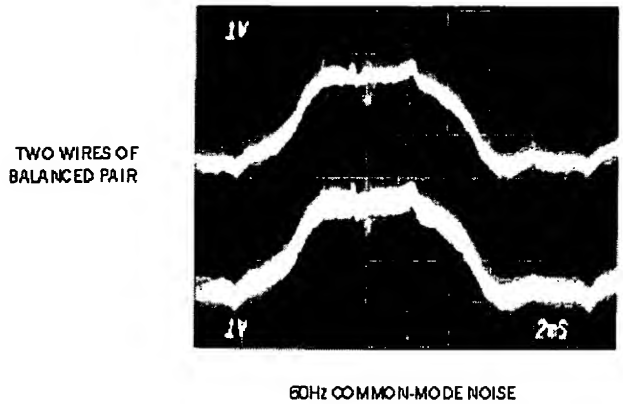


Figure 9. *Thanks to 60dB CMRR in the MAX436, the display in Figure 6 is unaffected when these 60Hz common-mode signals are deliberately added to each wire of the balanced cable.*

Although tests on the Figure 6 circuit involved only NTSC video signals, the circuit should provide comparable performance for PAL signals, which have a chroma carrier of 4.43MHz (vs. 3.58MHz).

Settling time measurements

Quick response and avoidable output saturation favor the MAX436 for use in measuring the settling time of slower amplifiers (**Figure 10**). In the test circuit, you configure the device under test (DUT) as a voltage follower and drive its inputs with a square wave. The MAX436 observes DUT settling time by comparing its input and output signals.

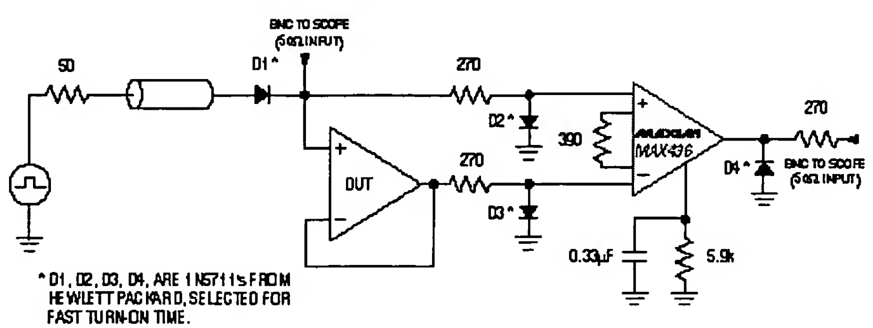


Figure 10. *Wideband differential inputs and an absence of output saturation suit the MAX436 for use in settling-time fixtures.*

The applied square wave appears quickly at the MAX436's non-inverting input, but is delayed by propagation time through the DUT before reaching the inverting input. The result is a brief but high-amplitude signal (clamped by D2 and D3) that appears between the MAX436 inputs

before the DUT can settle. If the MAX436 were a voltage-mode amplifier, this large differential input would cause the output transistors to saturate, thereby corrupting the settling-time measurement with overload-recovery time.

With properly chosen gain elements, however, the MAX436 can accommodate input signals that span its entire input common-mode range without saturation in the output stage. This characteristic suits the amplifier for settling-time measurements of D/A converters as well as high-speed op amps. (Following a 0.5V common-mode step, the MAX436 itself settles to ±0.1% in about 17ns.) Note that this common-mode response is faster than the response to a differential signal, in which the output response time is limited by the slew rate.

Figure 11 illustrates the response of a MAX442 (2-channel, 140MHz video multiplexer and amplifier) operating as a DUT in the circuit of Figure 6. The input step is 2V in this case. Note that the initial output level (40mV) should ideally be zero. It represents the difference in forward voltages for the Schottky clamp diodes D2 and D3, multiplied by voltage gain from the MAX436 to the scope (which is 8+50/390, i.e., near unity). This initial voltage has no effect on the settling measurement.

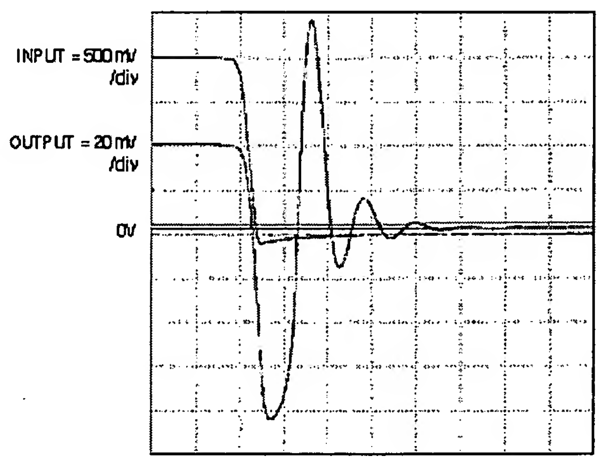


Figure 11. *Settling time for a MAX442 video amplifier in the Figure 10 circuit is 42ns.*

You can define settling time either from the beginning of the input's downward transition (which includes the DUT's propagation delay), or from the first output transition (a useful parameter in video applications). Because the MAX442's propagation delay is small, its ±0.1% settling time measures about 42ns either way. The mid-screen graticule line is 0V, the first cursor line is the final-settling level, and the next cursor line marks the boundary for ±0.1% settling.

VIDEO DESIGNERS: Join our E-mail-based Discussion Group for Analog Video Designers.

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MORE INFORMATION

MAX435: QuickView -- Full (PDF) Data Sheet (712k) -- Free Sample

MAX436: QuickView -- Full (PDF) Data Sheet (712k) -- Free Sample

*the **Technology Interface***

Application of the Operational Transconductance Amplifier (OTA) to Voltage-controlled Amplifiers and Active Filters

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Abstract

The application of the operational transconductance amplifier (OTA) in the design of simple amplifiers with voltage-controllable gain and to the design of firs-order and second-order active filters with controllable gains and controllable critical frequencies is demonstrated. A typical biasing scheme is also shown so that readers can more easily set up the circuits themselves.

Introduction

This paper demonstrates the usefulness of the operational transconductance amplifier (OTA) as a replacement for the conventional op-amp in both first and second-order active filters. It is at least partially intended to acquaint the technology student w ith the rudiments of operation of the OTA, as well as the practicalities of using the presently available commercial OTA's.

The structure of this paper is as follows. First, the basic operation of the OTA, including DC and AC operation, is explained. Simple example circuits will be presented in this section to demonstrate the similarities and differences between circuits w hich use the conventional op-amp and the OTA. Second, OTA active filter circuits will be presented and analyzed. The richness of the filter possibilities inherent in the second-order structures will be demonstrated. Finally, the last section will presen t practical considerations which must be considered when using the present generation of OTA's.

II. Basic OTA Operation

1. DC Operation

The OTA is a transconductance type device, which means that the input voltage controls an output current by means of the device transconductance, labeled g_m. This makes the OTA a voltage-controlled current source (VCCS), which is in contrast to the conventional op-amp, which is a voltage-controlled voltage source (VCVS). What is important and useful about the OTA's transconductance parameter is that it is controlled by an external current, the amplifier bias current, I_{ABC}, so that one obtains

$$g_m = \frac{I_{ABC}}{2V_T} = \frac{20}{V} \cdot I_{ABC} \quad \text{(Eq. 1)}$$

From this externally controlled transconductance, the output current as a function of the applied voltage difference between the two input pins, labeled v_+ and v_-, is given by

$$I_O = g_m(v_+ - v_-) \quad \text{(Eq. 2)}$$

Clearly, an output voltage can be derived from this current by simply driving a resistive load. The equivalent circuit for the OTA is shown in Figure 1.

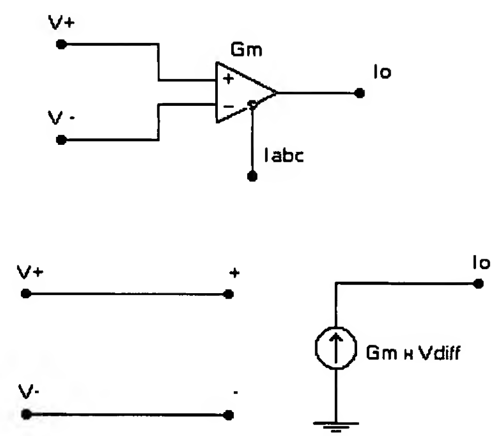

Fig. 1

At this point, two key differences between the OTA and the conventional op-amp must be kept in mind. First, since the OTA is a current source, the output impedance of the device is high, in contrast to the op-amp's very low output impedance. Because a low output impedance is often a desirable trait in general amplifiers used to drive resistive loads, certain of the newer commercial OTA's, such as National Semiconductor's LM13600, have on-chip controlled impedance buffers. Second, it is possible to desi gn circuits using the OTA that do not

employ negative feedback. In other words, instead of employing feedback to reduce the sensitivity of a circuit's performance to device parameters, the transconductance is treated as a design parameter, much as resisto rs and capacitors are treated in op-amp based circuits.

The biasing of the OTA's internal circuitry is such that the total quiescent supply current [Soclof-91] is given by $I_{SUPPLY} = 3I_{ABC}$. This seems to imply that the OTA can be used in micropower applications, even down to $I_{ABC} = 1\ \mu$ A. However, the losses in speed and bandwidth, which are controlled ultimately by I_{ABC}, can be severe at such low current levels.

2. AC Analysis and Frequency Response

Much of the dependence of open- and closed-loop bandwidth and frequency responses in the OTA are similar to those in the conventional op-amp. For a circuit employing negative feedback, a very important relationship between the closed-loop bandwidth, the amplifier bias current, and the closed-loop gain exists:

$$BW_{CL} = \left(\frac{20}{V}\right) I_{ABC} / [2\pi C_{NET} A_{CL}(0)] \quad \text{(Eq. 3)}$$

where C_{NET} is the sum of device junction capacitances at the output of the OTA and whatever load capacitance is attached to the circuit: $C_{NET} = C_O + C_L$. Equation (3) has the interesting consequence that cer tain types of active networks, such as active filters, can have their critical frequencies controlled by the external current, I_{ABC}, which of course can be in turn controlled by an external voltage.

III. OTA Circuits

1. Basic Voltage Amplifiers

This section will discuss a subset of general voltage amplifiers, both with and without negative feedback. For further information about the rich variety of amplifier configurations available using the OTA, the reference [Geiger-85] is extremely useful.. Figure 2 displays an inverting amplifier realized with an OTA which can provide not only controllable gain, but which also uses negative feedback to reduce the output resistance. Indeed, the output resistance is also now controllable via the transconduct ance. The voltage gain and output impedance are given by

$$\frac{V_O}{V_i} = \frac{1 - g_m R_2}{1 + g_m R_1} \quad \text{(Eq. 4)}$$

$$z_O = \frac{R_1 + R_2}{1 + g_m R_1} \quad \text{(Eq. 5)}$$

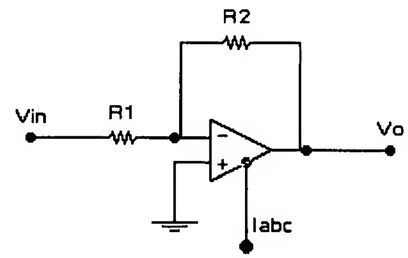

Fig. 2

The derivation of Equation (4) is presented in an appendix in order to demonstrate the typical analysis needed when dealing with OTA-based circuits. For the circuit in Figure 2, Equations (4) and (5) reduce to approximate forms for the case in which $g_m R_1 >> 1$; in this case, one obtains

$$\frac{V_O}{V_i} \cong \frac{-R_2}{R_1} \quad \text{(Eq. 6)}$$

$$Z_O \cong \frac{R_1 + R_2}{g_m R_1} \quad \text{(Eq. 7)}$$

Equation (6), of course, is nothing else than the in of a non-inverting amplifier. This is to be expected, since one of the properties of negative feedback is the nearly complete dependence of the gain on the feedback ratio only.

The final example of a basic building block amplifier using OTA's is shown in Figure 3. This is an example of an all-OTA amplifier, with the voltage gain and output impedance given by

$$\frac{V_O}{V_i} = -\frac{g_{m1}}{g_{m2}} \quad \text{(Eq. 8)}$$

$$Z_O = 1/g_{m2} \quad \text{(Eq. 9)}$$

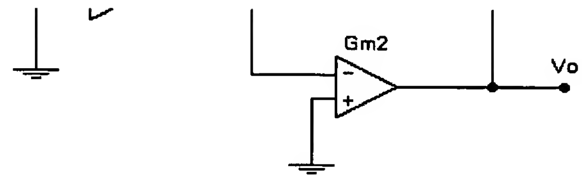

Fig. 3

The gain and output impedance are completely settable by the external currents, with no external, passive components except those needed to generate the current from a standard voltage source.

2. Active Filters with the OTA

Active filters are a standard application of the op-amp which can benefit greatly from the controllability of the OTA [Geiger-85]. The rudimentary theory and the circuit schematics for the basic active filters using op-amps are presented in many of th e textbooks in use in Electronics Technology or Engineering programs [Floyd-96 and Sedra-91]. What makes the OTA so attractive in these circuits is the ability to form filter circuits with voltage-variable control (via the I_{ABC} input) over a n umber of key performance parameters of the filter. The controlled parameter can be the midband gain of the circuit, as already realized in the simple circuits in the previous section. Alternatively, OTA-based active filters can use the external bias setti ng to control the location of the critical frequency, or 3-dB frequency, in a filter. The next logical step in controllability is the provision for independent gain and critical frequency setting. A number of other active filters can be realized with th e OTA. These provide the ability to not only change the critical frequency, the gain, or both, but also to preserve the shape of the response. For instance, one might want to control the critical frequency of the filter, but without altering the passban d ripple. It is even possible to change the type of response from lowpass to allpass to highpass by continuous adjustment of the transconductance g_m. Only a very few of these alternatives can be developed in this paper.

A very simple example of a first-order (one pole corresponding to a roll-off rate of -20 dB/decade in frequency) lowpass filter is shown in Figure 4. The voltage gain over the whole frequency range , and the -3 dB frequency, is given by

$$\frac{V_O}{V_i} = \frac{g_m}{sC + g_m} \quad \text{(Eq. 10)}$$

$$f_{3dB} = \frac{g_m}{2\pi C} \quad \text{(Eq. 11)}$$

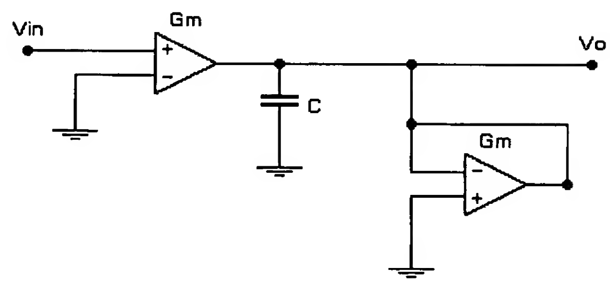

Fig. 4

In this circuit the second OTA, labeled "Gm2", is configured as a voltage variable resistor. It is this variable resistor which provides the variable cut-off frequency in Equation (11).

Figure 5 shows a second-order filter circuit with three voltage control terminals. Depending on which two of the three terminals are set to ground, one can realize a lowpass, highpass, bandpass, or notch filter. Each of these filters has a critical or center frequency which can be set by varying the transconductance, g_m, of the two OTA's in the circuit. These filters are called adjustable frequency constant-Q filters because they preserve the value of Q while the critical frequencies are s hifted. The derivation of the general relationship between the output voltage and the three control voltages is simply obtained:

$$I_{01} = g_{m1}\left(V_1^+ - V_1^-\right) = g_{m1}\left(V_A - V_{01}\right)$$

$$V_{C1} = I_{01} X_{C1} + V_B = V_2^+ = \frac{I_{01}}{sC_1} + V_B$$

$$I_{02} = g_{m2}\left(V_2^+ - V_2^-\right) = g_{m2}\left(\left(\frac{I_{01}}{sC_1} + V_B\right) - V_{01}\right)$$

$V_{01} = \dfrac{I_{02}}{sC_2} + V_C$. Upon substituting I_{02} and I_{01} from above, one obtains

$V_{01} = \dfrac{g_{m1} g_{m2}\left(V_A - V_{01}\right)}{s^2 C_1 C_2} + \dfrac{g_{m2}}{sC_2}\left(V_B - V_{01}\right) + V_C$. Bringing all terms in V_{01} together and manipulating, one finally obtains the transfer relation:

$$V_{01} = \frac{g_{m1}g_{m2}V_A + sC_1g_{m2}V_B + s^2C_1C_2V_C}{s^2C_1C_2 + sC_1g_{m2} + g_{m1}g_{m2}} \quad \text{(Eq. 12)}$$

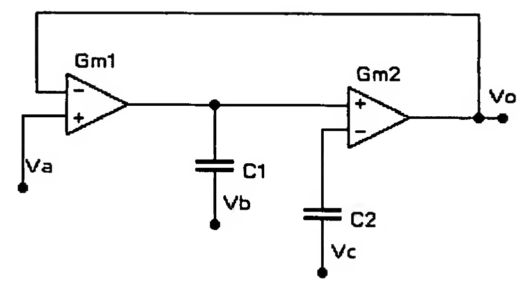

Fig. 5

In the above expressions, s is the complex frequency, $s = j\omega$, and I_{01} and I_{02} are the output currents for the first and second OTA's, respectively.

An example of the reduction of Equation (12) to a specific filter type is provided by making the following settings:

- Set $V_{in} = V_A$; V_B and V_C are grounded.
- Set $g_{m1} = g_{m2} = g_m$.
- Divide through by C_1C_2 in both numerator and denominator to achieve a standard biquadratic form.

The result is the following transfer function:

$$\frac{V_{01}}{V_A} = \frac{g_m^2 / C_1C_2}{s^2 + \frac{sg_m}{C_2} + \frac{g_m^2}{C_1C_2}} \quad \text{(Eq. 13)}$$

This expression has the form of the standard biquadratic circuit [Sedra-91]:

$$\frac{V_{01}(s)}{V_A(s)} = \frac{\omega_0^2}{s^2 + s\left(\frac{\omega_0}{Q}\right) + \omega_0^2} \quad \text{(Eq. 14)}$$

Therefore, the circuit with these particular control voltage settings is a lowpass filter with a critical frequency given by :

$$f_0 = \frac{g_m}{2\pi\sqrt{C_1 C_2}} \quad \text{(Eq. 15)}$$

and a constant $Q = \sqrt{\frac{C_2}{C_1}}$. It is straightforward to show that the following transfer functions can be obtained from the indicated control voltage settings:

- $V_{in} = V_B$; V_A and V_C grounded ⇒ Bandpass filter.
- $V_{in} = V_C$; V_A and V_B grounded ⇒ Highpass filter.
- $V_{in} = V_A = V_C$; V_B grounded ⇒ Notch filter.

3. Some Non-ideal Features of the OTA

One of the biggest drawbacks of the first versions [Harris-96, National-95] of the OTA was the limited range of the input differential voltage swing. This statement must be qualified in at least two ways. First, the limited input voltage swing appli es only if the OTA is being used in the open-loop configuration. In that case, if the difference-mode voltage exceeds about 25 mV, and the load resistance is relatively low (so that the open-loop gain is relatively small), then the circuit is no longer o perating in the linear region. This results in the output signal being distorted because of a nonlinear voltage transfer function [Sedra-91]. Of course, for circuits which make use of negative feedback, i.e., are operated in closed-loop conditions, then linear behavior is maintained.

The second qualifying remark is that the more recent versions of the OTA, such as the Harris CA3280A, National Semiconductor's LM13600, and Philips' NE5517, all use internal linearizing diodes at the input differential pair of the OTA. These make th e OTA's output current a linear function of the amplifier bias current over a wide range of differential input voltages. Figure 6 displays a typical biasing scheme for a generic commercial OTA. The control voltage, V_{CTL}, is used to generate t he amplifier bias current, I_{ABC}, through the resistor

R_{ABC}. The linearizing diodes, which are incorporated on-chip in the commercial OTA's mentioned above, can be biased on through the positive power supply voltage, $+V_{CC}$. The analysis of the linearization circuit and its effect on the output current can be found in [Soclof-91] and in several of the application notes from the major vendors [National-95,Philips-94].

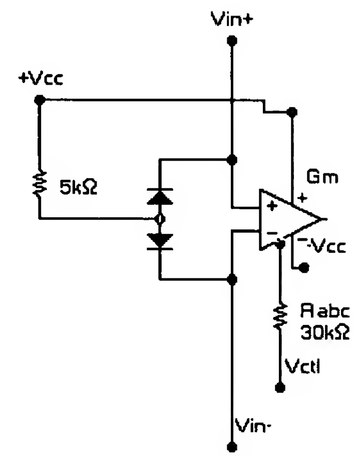

Fig. 6

IV. Conclusion

In conclusion, this paper has shown how the operational transconductance amplifier (OTA) adds controllability to a number of circuits commonly implemented with the conventional op-amp. In particular, an introduction has been given to the imp ortant class of voltage-controlled active filters realized with OTA's. Sufficient introductory material of a tutorial nature has been included that the technology student and instructor can make use of the paper for further investigations.

Appendix

Template for calculating with OTA-based circuits

This appendix is meant to demonstrate, for students in particular, how one uses the ideal OTA equivalent circuit in order to predict the behavior of simple OTA-based circuits in the lab. Some of the approach will be suggestive of approaches taken t o the standard op-amp. In particular, the ideal OTA, just like the ideal op-amp, has a nearly infinite input resistance. This means that no current enters the OTA input

pins, either inverting or non-inverting. However, the OTA is used often in open-loop, and therefore it is wise to learn how to treat the two input pins independently, as a virtual short circuit can not be assured in many configurations.

The circuit derivation that will be presented here is of the transfer function of, the inverting amplifier, Figure 2. The current I_{ABC} is presumed to be derived from a suitable biasing circuit. From the basic behavior of the OTA as a voltage controlled current source, one obtains

$$I_O = g_m\left(V^+ - V^-\right) = -g_m V^-$$

KVL around the OTA gives

$$\frac{V_i - V_-}{R_1} = I_1 = I_2 = \frac{V_- - V_O}{R_2}$$

But $I_O = -I_1 = -I_2$

Setting first $I_O = -I_2$ one then obtains

$$\frac{V_- - V_O}{R_2} = -(-g_m V_-) = g_m V_-$$. Gathering terms

$V_-(1 - g_m R_2) = V_O$. This gives

$V_- = \dfrac{V_O}{1 - g_m R_2}$. Now we can eliminate V_- by setting $I_2 = I_1$ and substituting for V_-:

$\dfrac{V_i}{R_1} - \dfrac{1}{R_1}\left(\dfrac{V_O}{1 - g_m R_2}\right) = \dfrac{1}{R_2}\left(\dfrac{V_O}{1 - g_m R_2}\right) - \dfrac{V_O}{R_2}$. Gathering all terms in V_O on the right - hand side

$V_i = V_O\left[\dfrac{1}{1 - g_m R_2} + \dfrac{R_1}{R_2}\left(\dfrac{g_m R_2}{1 - g_m R_2}\right)\right] = V_O\left(\dfrac{1 + g_m R_1}{1 - g_m R_2}\right)$. From this last expression, one obtains the voltage gain

$$\frac{V_O}{V_i} = \frac{1 - g_m R_2}{1 + g_m R_1}$$

This completes the derivation of the voltage gain for this amplifier. Repeated application of these same assumptions and operations will yield the corresponding expressions for the other circuits discussed in the main body of the text.< /P>

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National Semiconductor

June 2004

LM13700
Dual Operational Transconductance Amplifiers with Linearizing Diodes and Buffers

General Description

The LM13700 series consists of two current controlled transconductance amplifiers, each with differential inputs and a push-pull output. The two amplifiers share common supplies but otherwise operate independently. Linearizing diodes are provided at the inputs to reduce distortion and allow higher input levels. The result is a 10 dB signal-to-noise improvement referenced to 0.5 percent THD. High impedance buffers are provided which are especially designed to complement the dynamic range of the amplifiers. The output buffers of the LM13700 differ from those of the LM13600 in that their input bias currents (and hence their output DC levels) are independent of I_{ABC}. This may result in performance superior to that of the LM13600 in audio applications.

Features

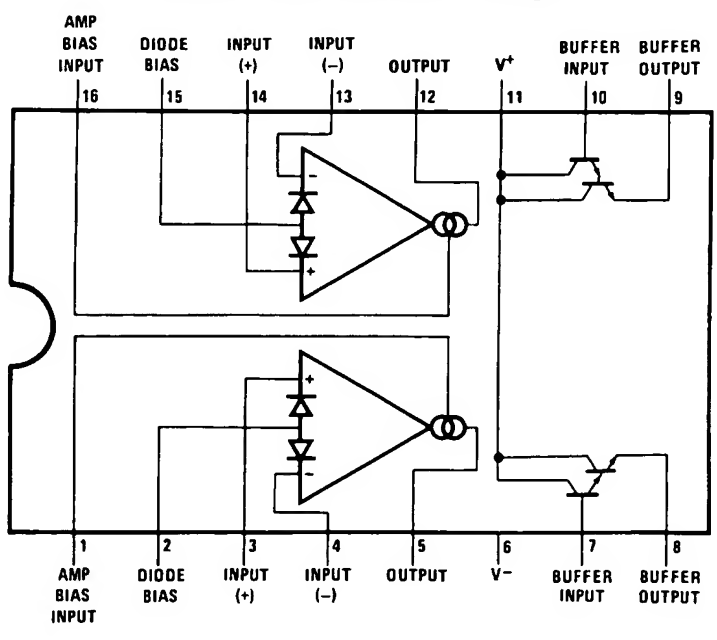
- g_m adjustable over 6 decades
- Excellent g_m linearity
- Excellent matching between amplifiers
- Linearizing diodes
- High impedance buffers
- High output signal-to-noise ratio

Applications

- Current-controlled amplifiers
- Current-controlled impedances
- Current-controlled filters
- Current-controlled oscillators
- Multiplexers
- Timers
- Sample-and-hold circuits

Connection Diagram

Dual-In-Line and Small Outline Packages



Top View
Order Number LM13700M, LM13700MX or LM13700N
See NS Package Number M16A or N16A

 DS007981 www.national.com

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

Supply Voltage	
LM13700	36 V_{DC} or ±18V
Power Dissipation (Note 2) T_A = 25°C	
LM13700N	570 mW
Differential Input Voltage	±5V
Diode Bias Current (I_D)	2 mA
Amplifier Bias Current (I_{ABC})	2 mA
Output Short Circuit Duration	Continuous
Buffer Output Current (Note 3)	20 mA
Operating Temperature Range	
LM13700N	0°C to +70°C
DC Input Voltage	$+V_S$ to $-V_S$
Storage Temperature Range	−65°C to +150°C
Soldering Information	
Dual-In-Line Package	
Soldering (10 sec.)	260°C
Small Outline Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

Electrical Characteristics (Note 4)

Parameter	Conditions	LM13700 Min	Typ	Max	Units
Input Offset Voltage (V_{OS})	Over Specified Temperature Range		0.4	4	mV
	I_{ABC} = 5 µA		0.3	4	
V_{OS} Including Diodes	Diode Bias Current (I_D) = 500 µA		0.5	5	mV
Input Offset Change	5 µA ≤ I_{ABC} ≤ 500 µA		0.1	3	mV
Input Offset Current			0.1	0.6	µA
Input Bias Current	Over Specified Temperature Range		0.4	5	µA
			1	8	
Forward		6700	9600	13000	µmho
Transconductance (g_m)	Over Specified Temperature Range	5400			
g_m Tracking			0.3		dB
Peak Output Current	R_L = 0, I_{ABC} = 5 µA		5		µA
	R_L = 0, I_{ABC} = 500 µA	350	500	650	
	R_L = 0, Over Specified Temp Range	300			
Peak Output Voltage					
Positive	R_L = ∞, 5 µA ≤ I_{ABC} ≤ 500 µA	+12	+14.2		V
Negative	R_L = ∞, 5 µA ≤ I_{ABC} ≤ 500 µA	−12	−14.4		V
Supply Current	I_{ABC} = 500 µA, Both Channels		2.6		mA
V_{OS} Sensitivity					
Positive	$\Delta V_{OS}/\Delta V^+$		20	150	µV/V
Negative	$\Delta V_{OS}/\Delta V^-$		20	150	µV/V
CMRR		80	110		dB
Common Mode Range		±12	±13.5		V
Crosstalk	Referred to Input (Note 5) 20 Hz < f < 20 kHz		100		dB
Differential Input Current	I_{ABC} = 0, Input = ±4V		0.02	100	nA
Leakage Current	I_{ABC} = 0 (Refer to Test Circuit)		0.2	100	nA
Input Resistance		10	26		kΩ
Open Loop Bandwidth			2		MHz
Slew Rate	Unity Gain Compensated		50		V/µs
Buffer Input Current	(Note 5)		0.5	2	µA
Peak Buffer Output Voltage	(Note 5)	10			V

Note 1: "Absolute Maximum Ratings" indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is functional, but do not guarantee specific performance limits.

Note 2: For operation at ambient temperatures above 25°C, the device must be derated based on a 150°C maximum junction temperature and a thermal resistance, junction to ambient, as follows: LM13700N, 90°C/W; LM13700M, 110°C/W.

Note 3: Buffer output current should be limited so as to not exceed package dissipation.

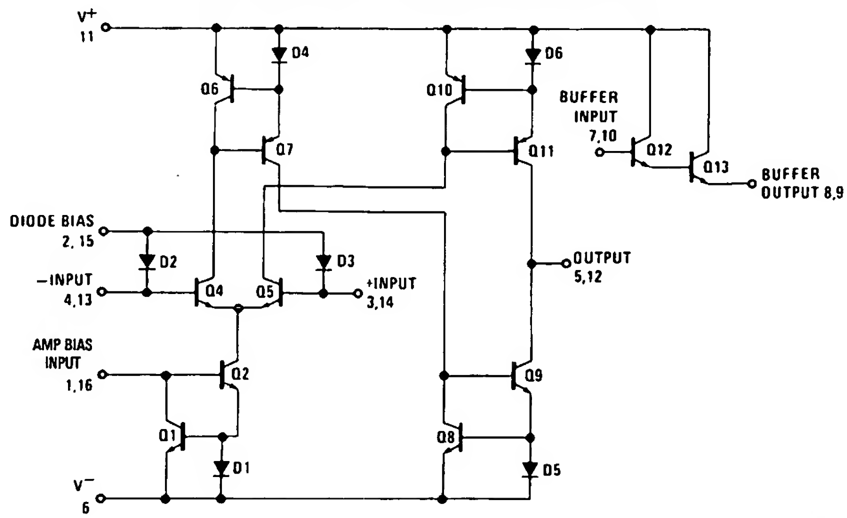
Electrical Characteristics (Note 4) (Continued)

Note 4: These specifications apply for $V_S = \pm 15V$, $T_A = 25°C$, amplifier bias current (I_{ABC}) = 500 µA, pins 2 and 15 open unless otherwise specified. The inputs to the buffers are grounded and outputs are open.

Note 5: These specifications apply for $V_S = \pm 15V$, $I_{ABC} = 500$ µA, $R_{OUT} = 5$ kΩ connected from the buffer output to $-V_S$ and the input of the buffer is connected to the transconductance amplifier output.

Schematic Diagram

One Operational Transconductance Amplifier



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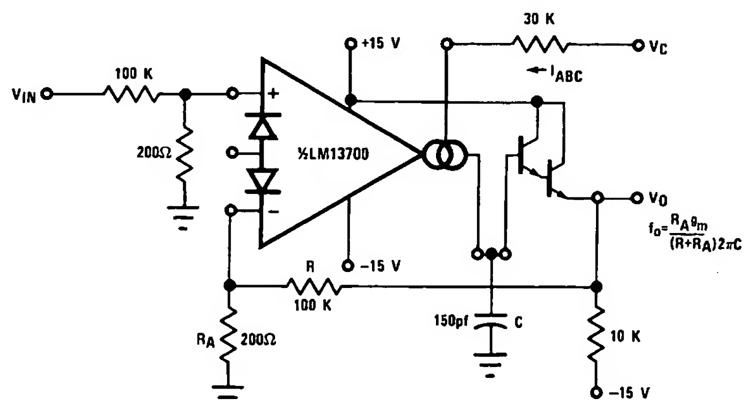
Typical Application



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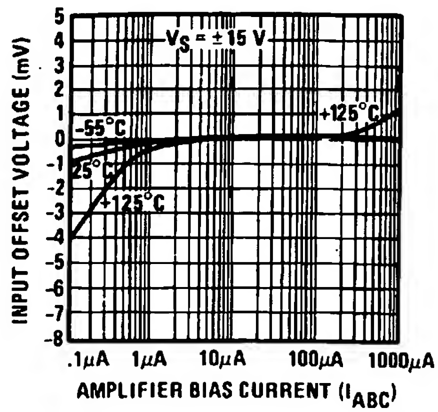
Voltage Controlled Low-Pass Filter

Typical Performance Characteristics

Input Offset Voltage



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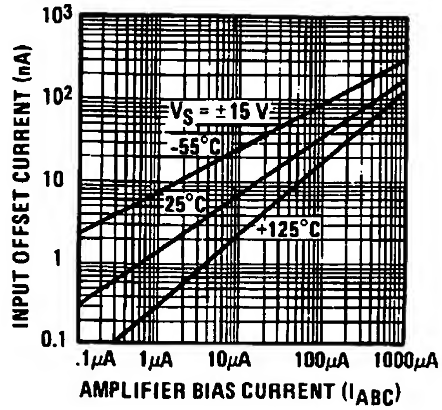
Input Offset Current



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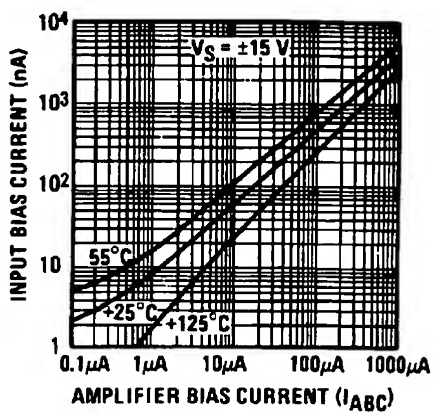
Input Bias Current



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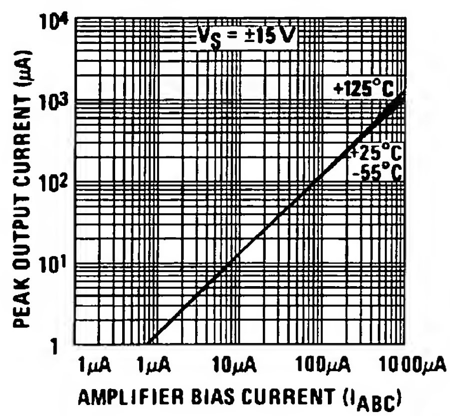
Peak Output Current



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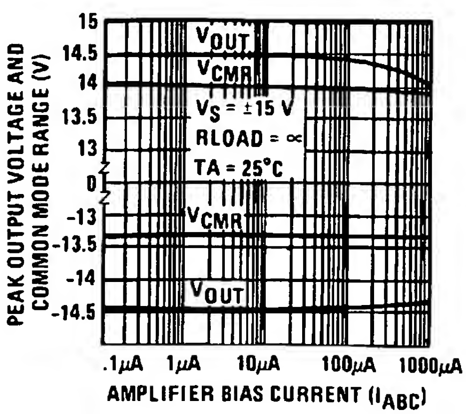
Peak Output Voltage and Common Mode Range



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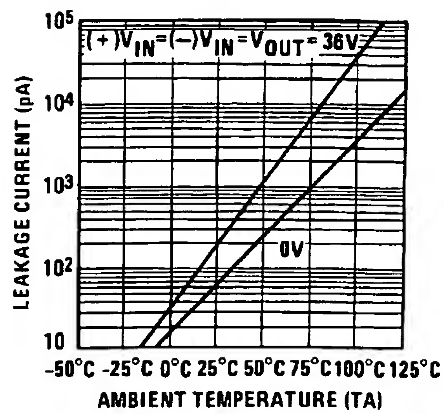
Leakage Current



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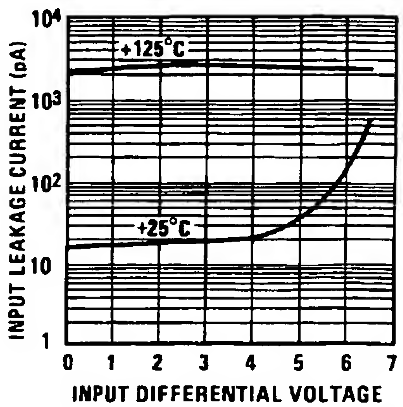
Typical Performance Characteristics (Continued)

Input Leakage



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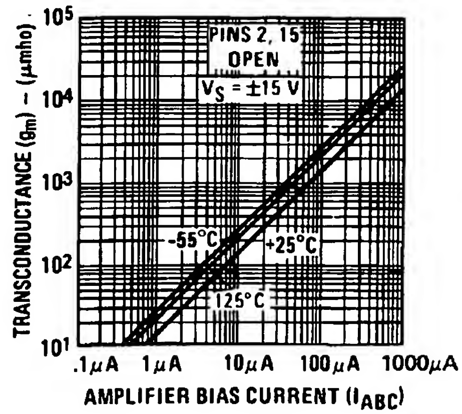
Transconductance



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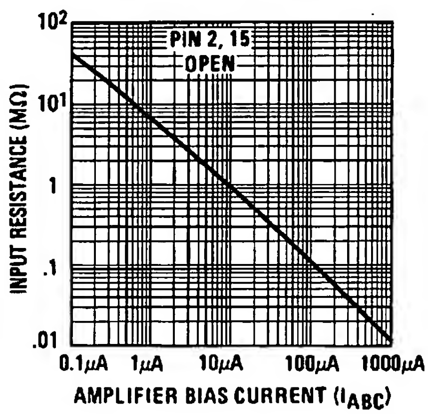
Input Resistance



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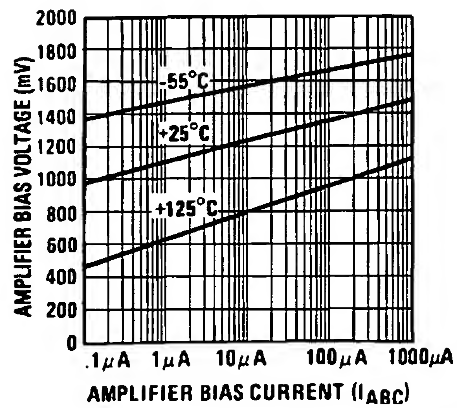
Amplifier Bias Voltage vs. Amplifier Bias Current



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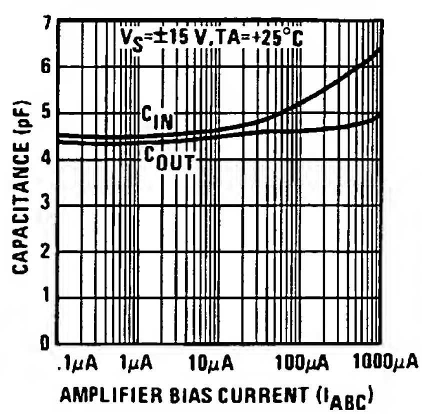
Input and Output Capacitance



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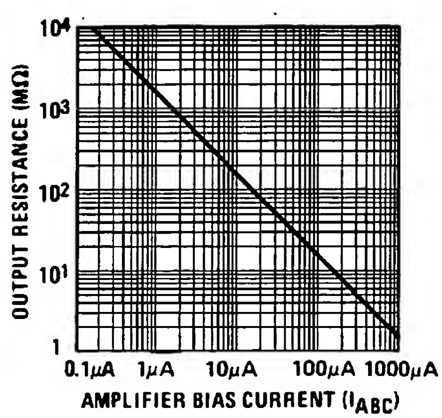
Output Resistance



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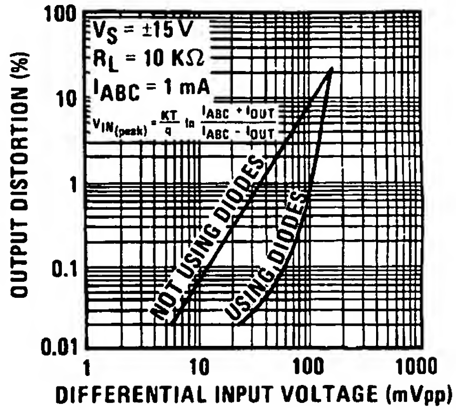
Typical Performance Characteristics (Continued)

Distortion vs. Differential Input Voltage



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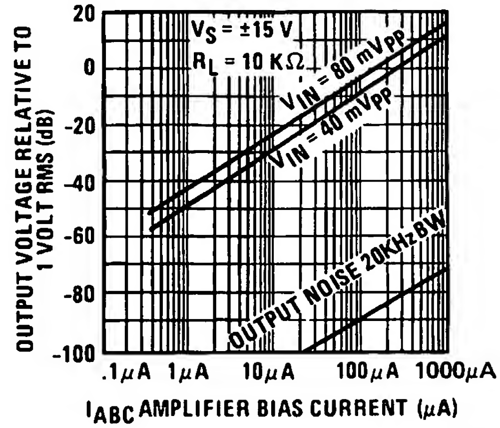
Voltage vs. Amplifier Bias Current



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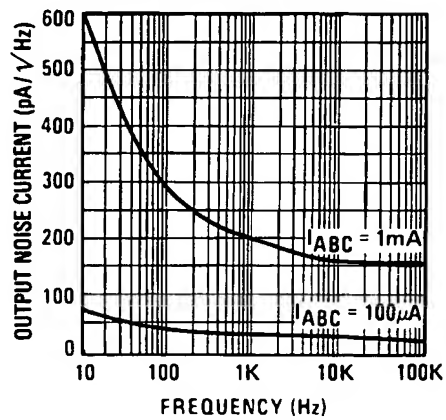
Output Noise vs. Frequency



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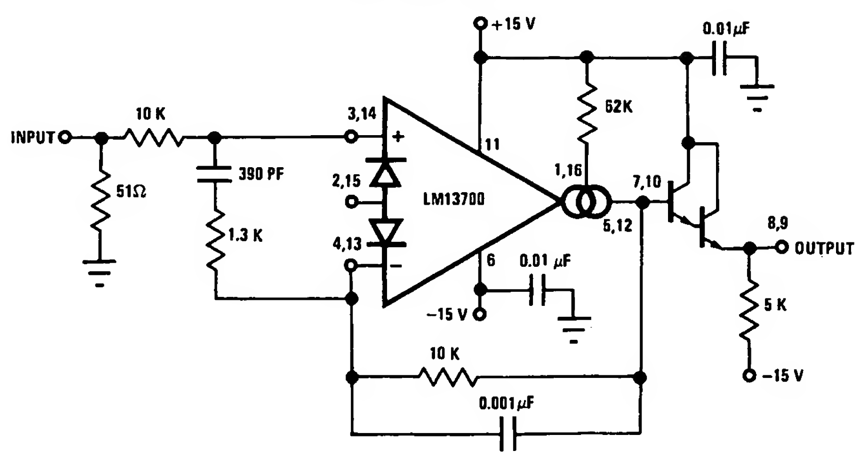
Unity Gain Follower



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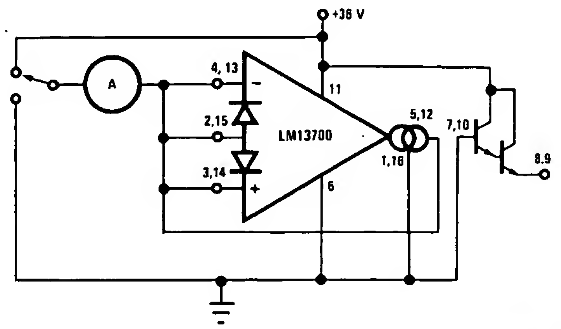
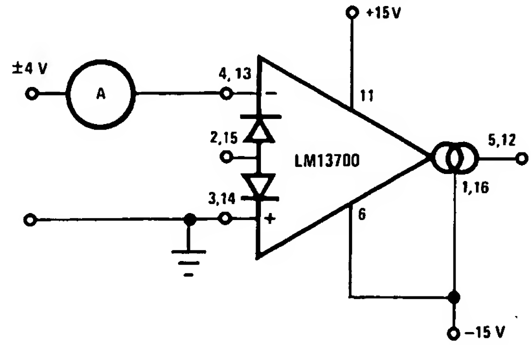
Typical Performance Characteristics (Continued)

Leakage Current Test Circuit



Differential Input Current Test Circuit



Circuit Description

The differential transistor pair Q_4 and Q_5 form a transconductance stage in that the ratio of their collector currents is defined by the differential input voltage according to the transfer function:

$$V_{IN} = \frac{kT}{q} \ln \frac{I_5}{I_4} \tag{1}$$

where V_{IN} is the differential input voltage, kT/q is approximately 26 mV at 25°C and I_5 and I_4 are the collector currents of transistors Q_5 and Q_4 respectively. With the exception of Q_{12} and Q_{13}, all transistors and diodes are identical in size. Transistors Q_1 and Q_2 with Diode D_1 form a current mirror which forces the sum of currents I_4 and I_5 to equal I_{ABC}:

$$I_4 + I_5 = I_{ABC} \tag{2}$$

where I_{ABC} is the amplifier bias current applied to the gain pin.

For small differential input voltages the ratio of I_4 and I_5 approaches unity and the Taylor series of the ln function can be approximated as:

$$\frac{kT}{q} \ln \frac{I_5}{I_4} \approx \frac{kT}{q} \frac{I_5 - I_4}{I_4}$$

$$I_4 \approx I_5 \approx \frac{I_{ABC}}{2} \tag{3}$$

$$V_{IN} \left[\frac{I_{ABC} q}{2kT}\right] = I_5 - I_4 \tag{4}$$

Collector currents I_4 and I_5 are not very useful by themselves and it is necessary to subtract one current from the other. The remaining transistors and diodes form three current mirrors that produce an output current equal to I_5 minus I_4 thus:

$$V_{IN} \left[\frac{I_{ABC} q}{2kT}\right] = I_{OUT} \tag{5}$$

The term in brackets is then the transconductance of the amplifier and is proportional to I_{ABC}.

Linearizing Diodes

For differential voltages greater than a few millivolts, *Equation (3)* becomes less valid and the transconductance becomes increasingly nonlinear. *Figure 1* demonstrates how the internal diodes can linearize the transfer function of the amplifier. For convenience assume the diodes are biased with current sources and the input signal is in the form of current I_S. Since the sum of I_4 and I_5 is I_{ABC} and the difference is I_{OUT}, currents I_4 and I_5 can be written as follows:

$$I_4 = \frac{I_{ABC}}{2} - \frac{I_{OUT}}{2},\ I_5 = \frac{I_{ABC}}{2} + \frac{I_{OUT}}{2}$$

Since the diodes and the input transistors have identical geometries and are subject to similar voltages and temperatures, the following is true:

$$\frac{kT}{q} \ln \frac{\frac{I_D}{2} + I_S}{\frac{I_D}{2} - I_S} = \frac{kT}{q} \ln \frac{\frac{I_{ABC}}{2} + \frac{I_{OUT}}{2}}{\frac{I_{ABC}}{2} - \frac{I_{OUT}}{2}}$$

$$\therefore I_{OUT} = I_S \left(\frac{2 I_{ABC}}{I_D}\right) \text{ for } |I_S| < \frac{I_D}{2} \tag{6}$$

Notice that in deriving *Equation (6)* no approximations have been made and there are no temperature-dependent terms. The limitations are that the signal current not exceed $I_D/2$ and that the diodes be biased with currents. In practice, replacing the current sources with resistors will generate insignificant errors.

Applications
Voltage Controlled Amplifiers

Figure 2 shows how the linearizing diodes can be used in a voltage-controlled amplifier. To understand the input biasing, it is best to consider the 13 kΩ resistor as a current source and use a Thevenin equivalent circuit as shown in *Figure 3*. This circuit is similar to *Figure 1* and operates the same. The potentiometer in *Figure 2* is adjusted to minimize the effects of the control signal at the output.

Applications

Voltage Controlled Amplifiers (Continued)

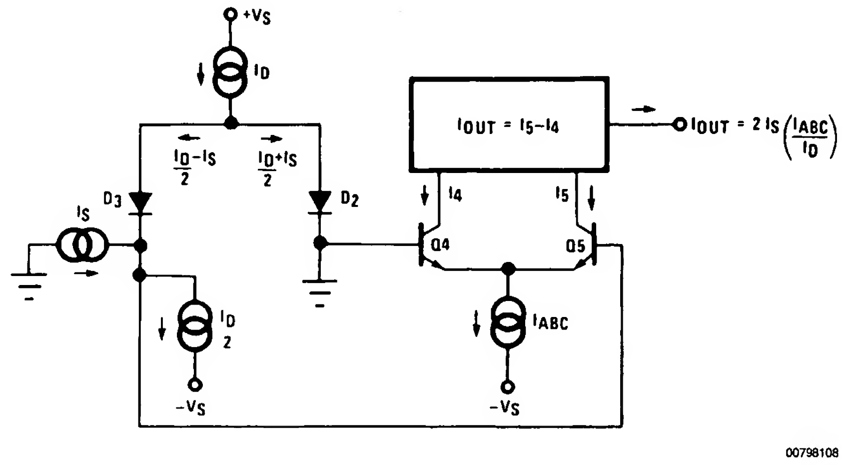


FIGURE 1. Linearizing Diodes

For optimum signal-to-noise performance, I_{ABC} should be as large as possible as shown by the Output Voltage vs. Amplifier Bias Current graph. Larger amplitudes of input signal also improve the S/N ratio. The linearizing diodes help here by allowing larger input signals for the same output distortion as shown by the Distortion vs. Differential Input Voltage graph. S/N may be optimized by adjusting the magnitude of the input signal via R_{IN} (*Figure 2*) until the output distortion is below some desired level. The output voltage swing can then be set at any level by selecting R_L.

Although the noise contribution of the linearizing diodes is negligible relative to the contribution of the amplifier's internal transistors, I_D should be as large as possible. This minimizes the dynamic junction resistance of the diodes (r_e) and maximizes their linearizing action when balanced against R_{IN}. A value of 1 mA is recommended for I_D unless the specific application demands otherwise.

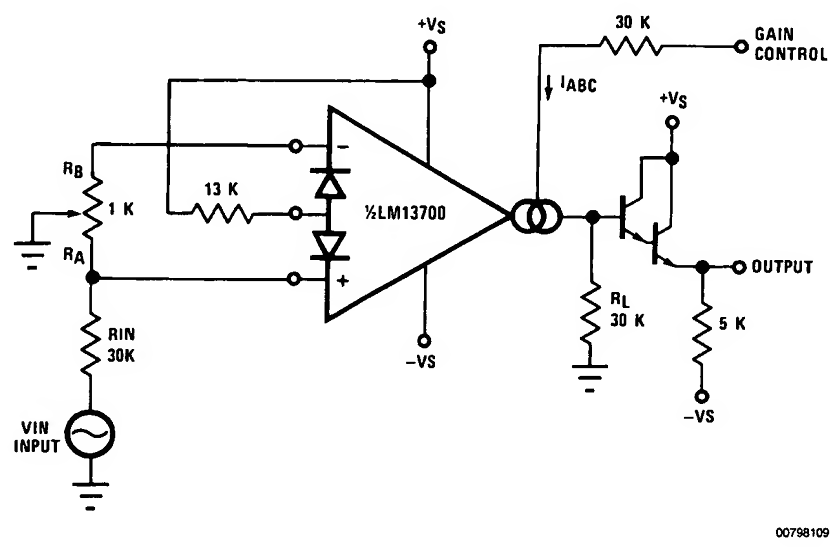


FIGURE 2. Voltage Controlled Amplifier

Applications
Voltage Controlled Amplifiers (Continued)

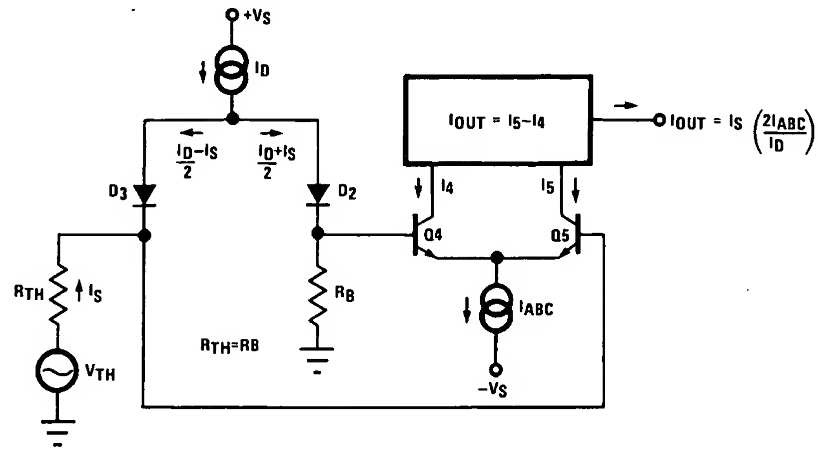


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FIGURE 3. Equivalent VCA Input Circuit

Stereo Volume Control

The circuit of *Figure 4* uses the excellent matching of the two LM13700 amplifiers to provide a Stereo Volume Control with a typical channel-to-channel gain tracking of 0.3 dB. R_P is provided to minimize the output offset voltage and may be replaced with two 510Ω resistors in AC-coupled applications. For the component values given, amplifier gain is derived for *Figure 2* as being:

$$\frac{V_O}{V_{IN}} = 940 \times I_{ABC}$$

If V_C is derived from a second signal source then the circuit becomes an amplitude modulator or two-quadrant multiplier as shown in *Figure 5*, where:

$$I_O = \frac{-2I_S}{I_D}(I_{ABC}) = \frac{-2I_S}{I_D}\frac{V_{IN2}}{R_C} - \frac{2I_S}{I_D}\frac{(V^- + 1.4V)}{R_C}$$

The constant term in the above equation may be cancelled by feeding I_S x $I_D R_C/2(V- + 1.4V)$ into I_O. The circuit of *Figure 6* adds R_M to provide this current, resulting in a four-quadrant multiplier where R_C is trimmed such that V_O = 0V for V_{IN2} = 0V. R_M also serves as the load resistor for I_O.

Stereo Volume Control (Continued)

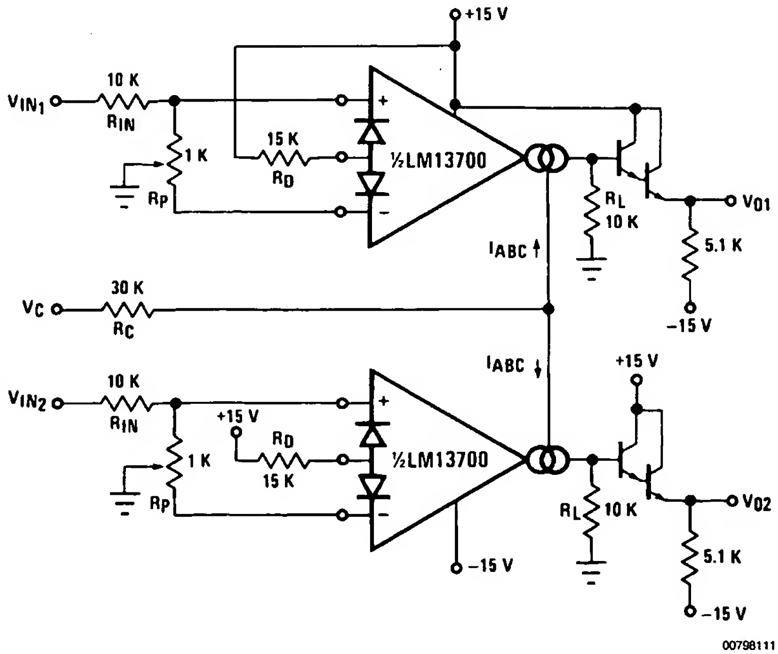


FIGURE 4. Stereo Volume Control

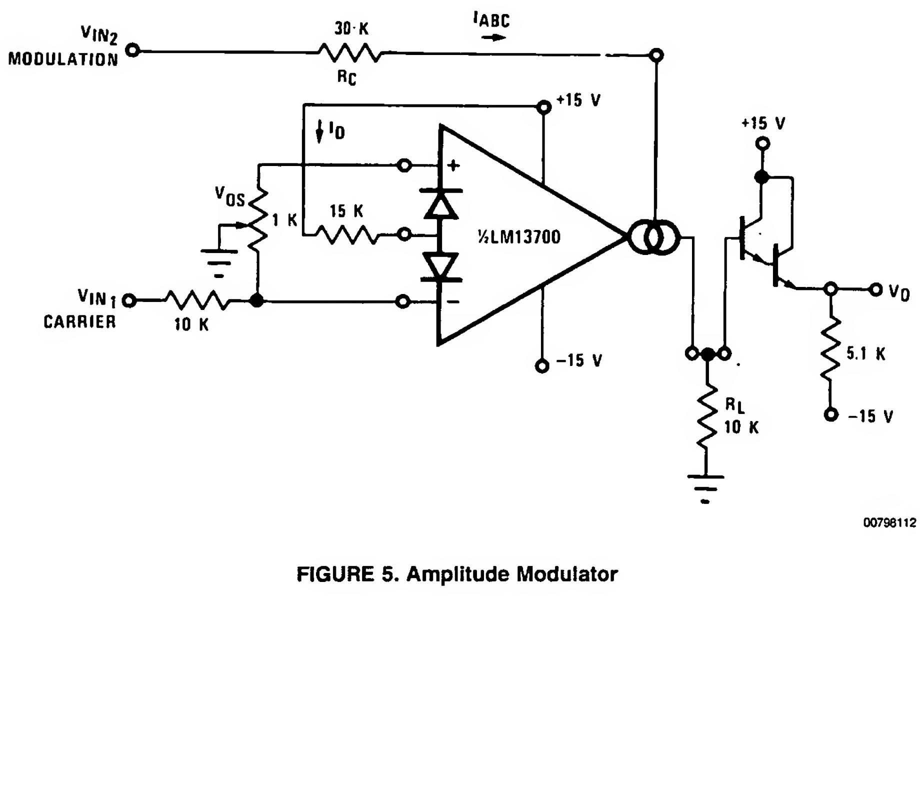


FIGURE 5. Amplitude Modulator

Stereo Volume Control (Continued)

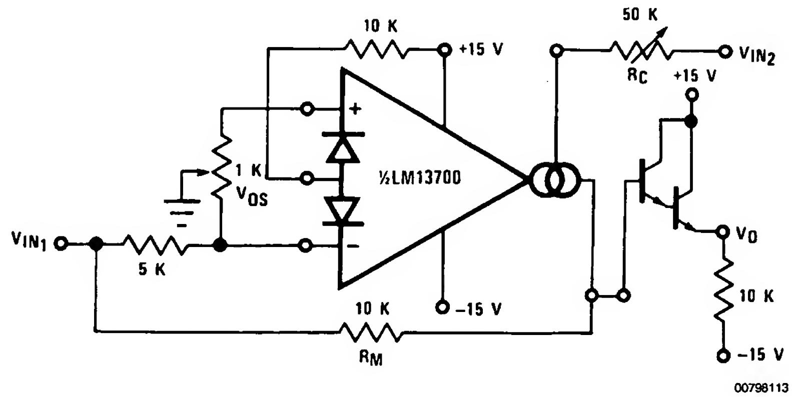


FIGURE 6. Four-Quadrant Multiplier

Noting that the gain of the LM13700 amplifier of *Figure 3* may be controlled by varying the linearizing diode current I_D as well as by varying I_{ABC}, *Figure 7* shows an AGC Amplifier using this approach. As V_O reaches a high enough amplitude ($3V_{BE}$) to turn on the Darlington transistors and the linearizing diodes, the increase in I_D reduces the amplifier gain so as to hold V_O at that level.

Voltage Controlled Resistors

An Operational Transconductance Amplifier (OTA) may be used to implement a Voltage Controlled Resistor as shown in *Figure 8*. A signal voltage applied at R_X generates a V_{IN} to the LM13700 which is then multiplied by the g_m of the amplifier to produce an output current, thus:

$$R_X = \frac{R + R_A}{g_m R_A}$$

where $g_m \approx 19.2 I_{ABC}$ at 25°C. Note that the attenuation of V_O by R and R_A is necessary to maintain V_{IN} within the linear range of the LM13700 input.

Figure 9 shows a similar VCR where the linearizing diodes are added, essentially improving the noise performance of the resistor. A floating VCR is shown in *Figure 10*, where each "end" of the "resistor" may be at any voltage within the output voltage range of the LM13700.

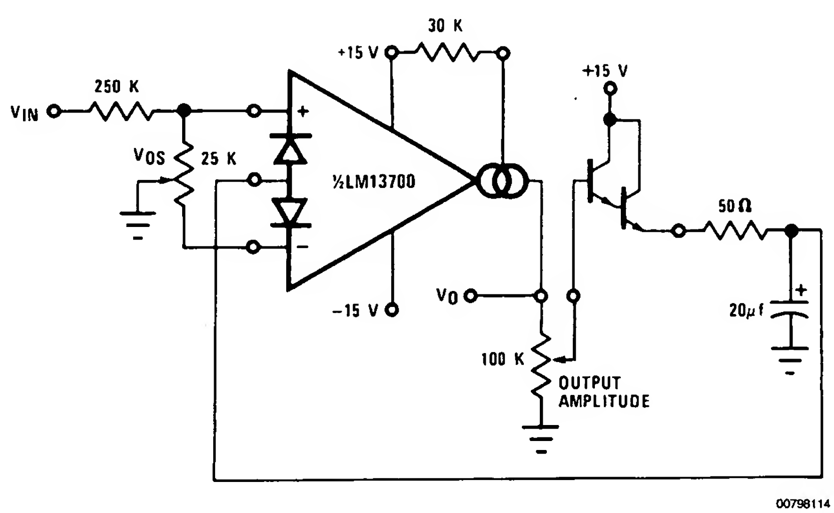


FIGURE 7. AGC Amplifier

Voltage Controlled Resistors (Continued)

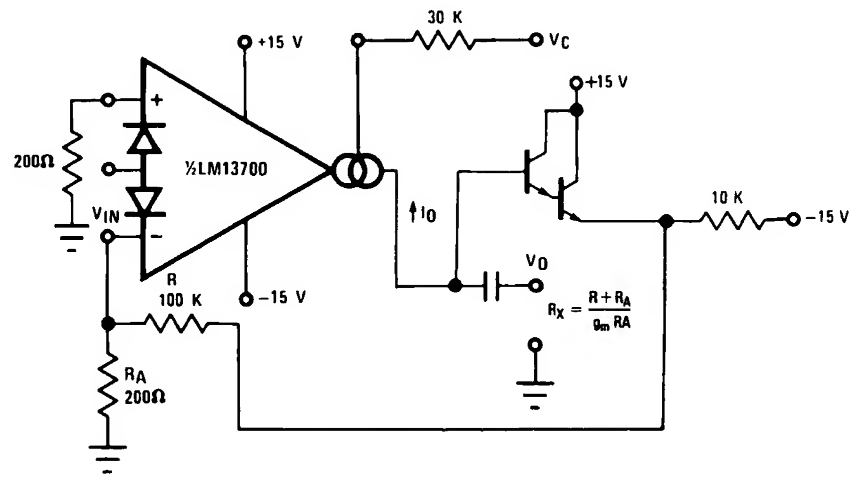


FIGURE 8. Voltage Controlled Resistor, Single-Ended

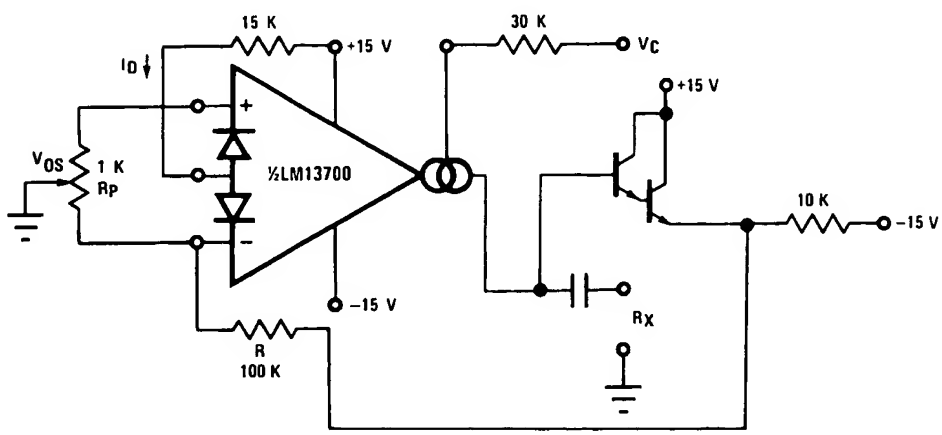


FIGURE 9. Voltage Controlled Resistor with Linearizing Diodes

Voltage Controlled Filters

OTA's are extremely useful for implementing voltage controlled filters, with the LM13700 having the advantage that the required buffers are included on the I.C. The VC Lo-Pass Filter of *Figure 11* performs as a unity-gain buffer amplifier at frequencies below cut-off, with the cut-off frequency being the point at which X_C/g_m equals the closed-loop gain of (R/R_A). At frequencies above cut-off the circuit provides a single RC roll-off (6 dB per octave) of the input signal amplitude with a −3 dB point defined by the given equation, where g_m is again 19.2 x I_{ABC} at room temperature. *Figure 12* shows a VC High-Pass Filter which operates in much the same manner, providing a single RC roll-off below the defined cut-off frequency.

Additional amplifiers may be used to implement higher order filters as demonstrated by the two-pole Butterworth Lo-Pass Filter of *Figure 13* and the state variable filter of *Figure 14*. Due to the excellent g_m tracking of the two amplifiers, these filters perform well over several decades of frequency.

Voltage Controlled Filters (Continued)

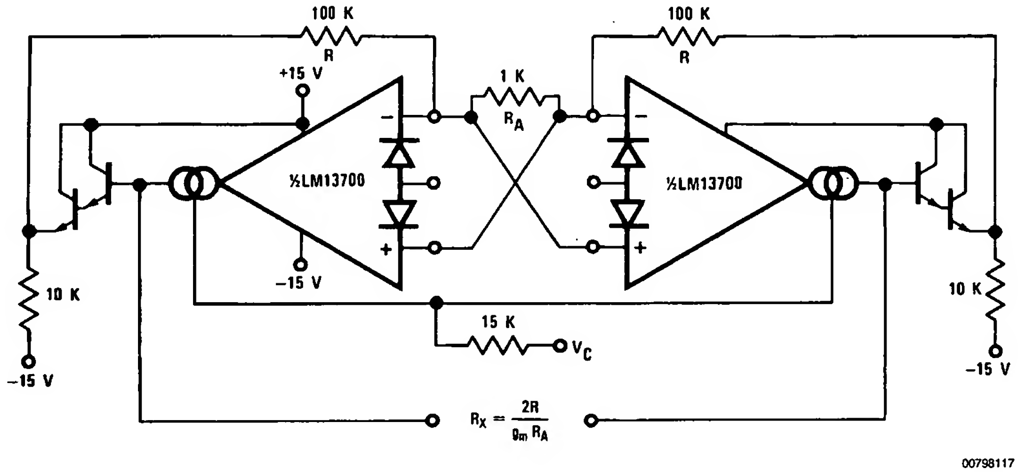


FIGURE 10. Floating Voltage Controlled Resistor

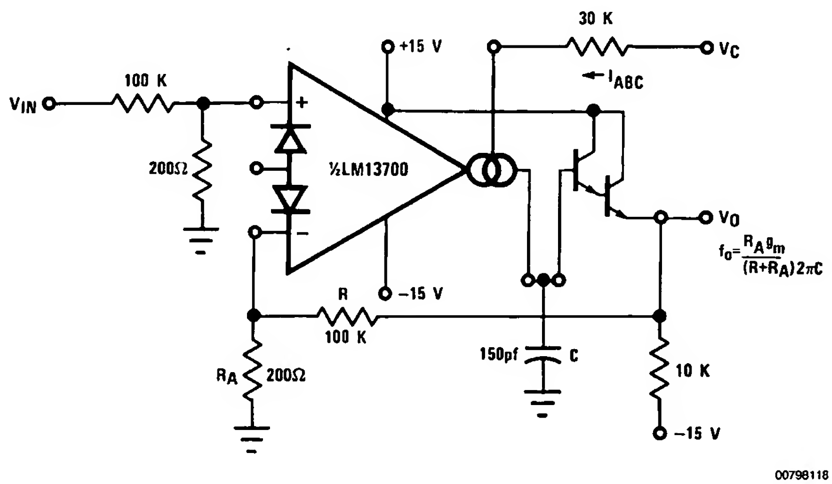
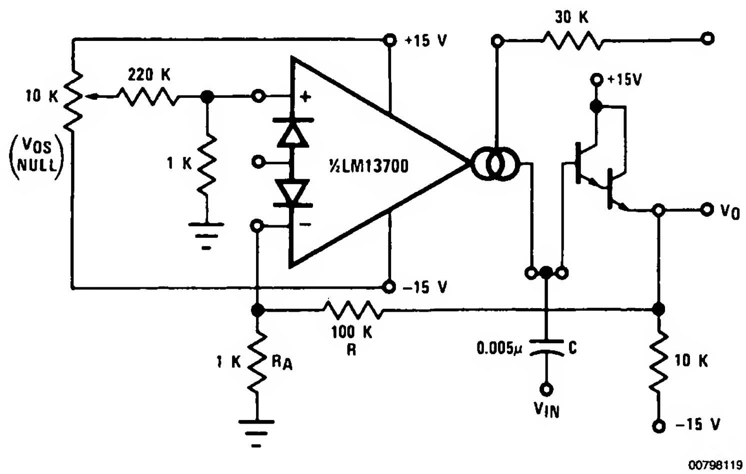


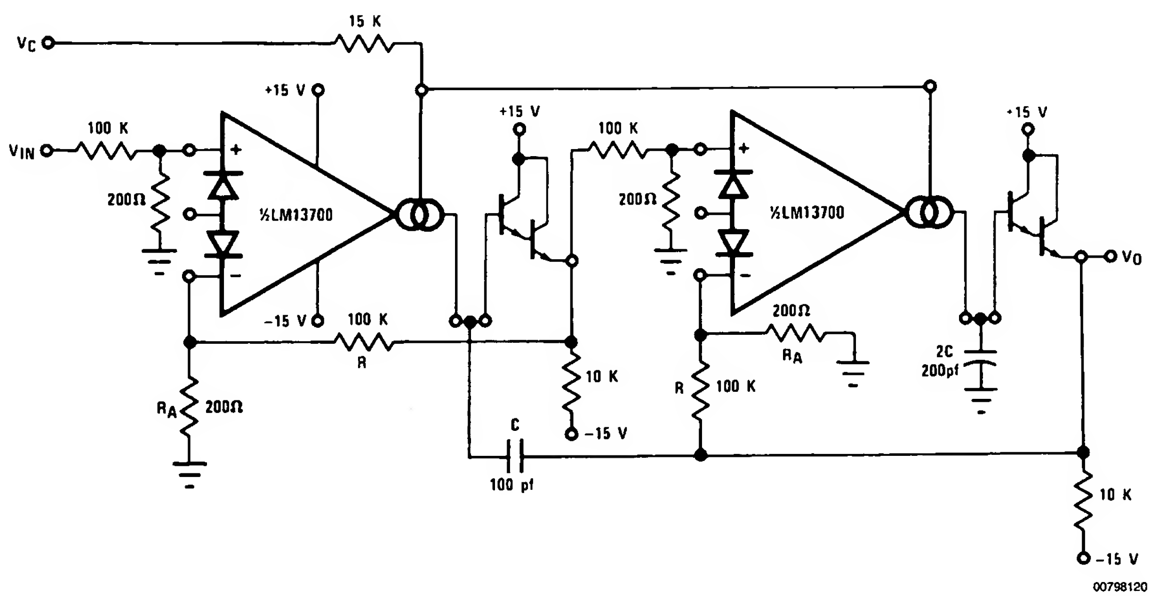
FIGURE 11. Voltage Controlled Low-Pass Filter

Voltage Controlled Filters (Continued)



$$f_o = \frac{R_A\, g_m}{(R + R_A)\, 2\pi C}$$

FIGURE 12. Voltage Controlled Hi-Pass Filter



$$f_o = \frac{R_A\, g_m}{(R + R_A)\, 2\pi C}$$

FIGURE 13. Voltage Controlled 2-Pole Butterworth Lo-Pass Filter

Voltage Controlled Filters (Continued)

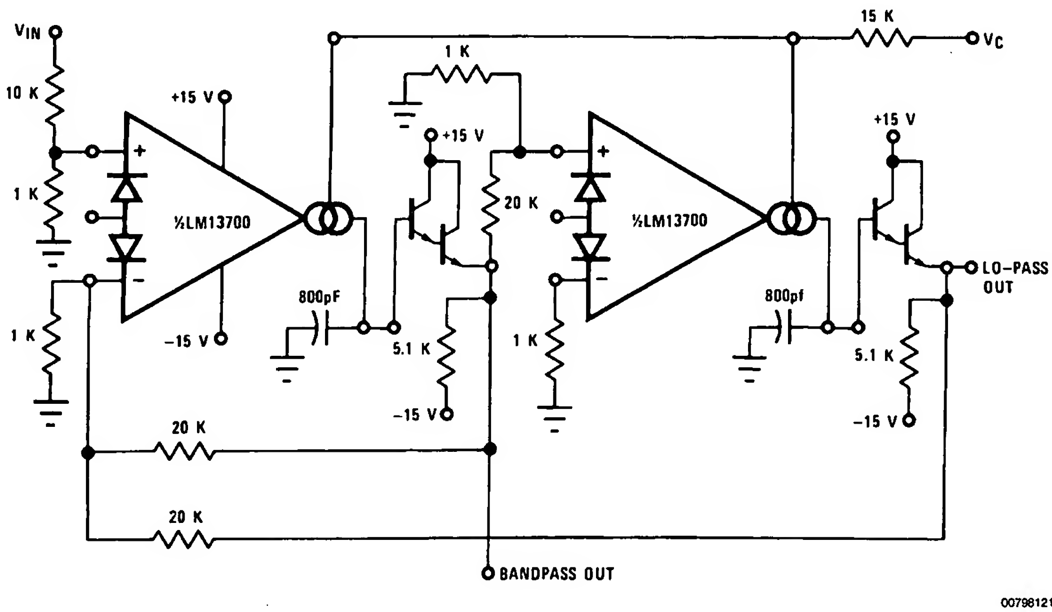


FIGURE 14. Voltage Controlled State Variable Filter

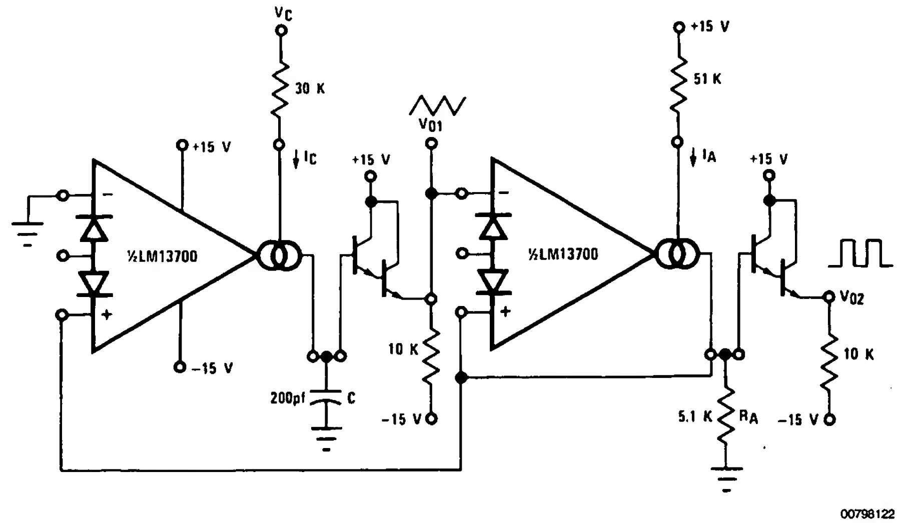
Voltage Controlled Oscillators

The classic Triangular/Square Wave VCO of *Figure 15* is one of a variety of Voltage Controlled Oscillators which may be built utilizing the LM13700. With the component values shown, this oscillator provides signals from 200 kHz to below 2 Hz as I_C is varied from 1 mA to 10 nA. The output amplitudes are set by I_A x R_A. Note that the peak differential input voltage must be less than 5V to prevent zenering the inputs.

A few modifications to this circuit produce the ramp/pulse VCO of *Figure 16*. When V_{O2} is high, I_F is added to I_C to increase amplifier A1's bias current and thus to increase the charging rate of capacitor C. When V_{O2} is low, I_F goes to zero and the capacitor discharge current is set by I_C.

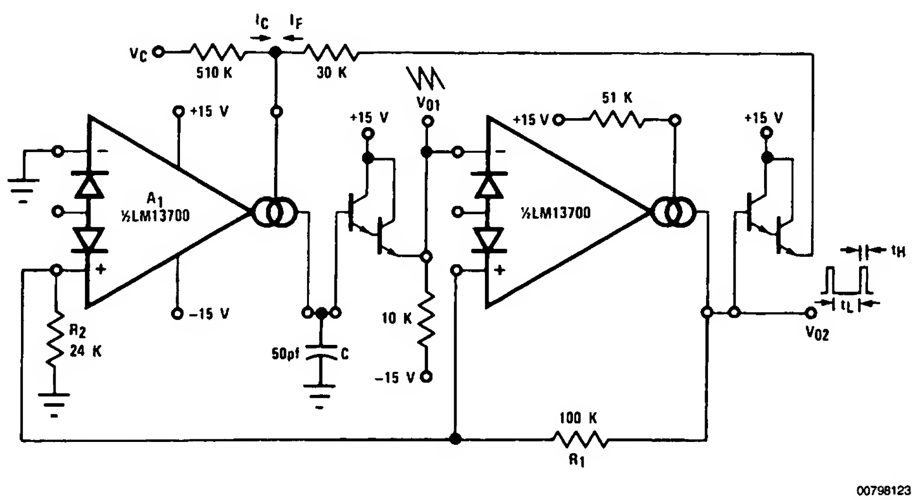
The VC Lo-Pass Filter of *Figure 11* may be used to produce a high-quality sinusoidal VCO. The circuit of *Figure 16* employs two LM13700 packages, with three of the amplifiers configured as lo-pass filters and the fourth as a limiter/inverter. The circuit oscillates at the frequency at which the loop phase-shift is 360° or 180° for the inverter and 60° per filter stage. This VCO operates from 5 Hz to 50 kHz with less than 1% THD.

Voltage Controlled Oscillators (Continued)



$$f_{OSC} = \frac{I_C}{4CI_AR_A}$$

FIGURE 15. Triangular/Square-Wave VCO



$$V_{PK} = \frac{(V^{+} \pm 0.8V)\,R_2}{R_1 + R_2}$$

$$t_H \approx \frac{2V_{PK}C}{I_F}$$

$$t_L = \frac{2V_{PK}C}{I_C}$$

$$f_0 \approx \frac{I_C}{2V_{PK}C} \text{ for } I_C << I_F$$

FIGURE 16. Ramp/Pulse VCO

Voltage Controlled Oscillators (Continued)

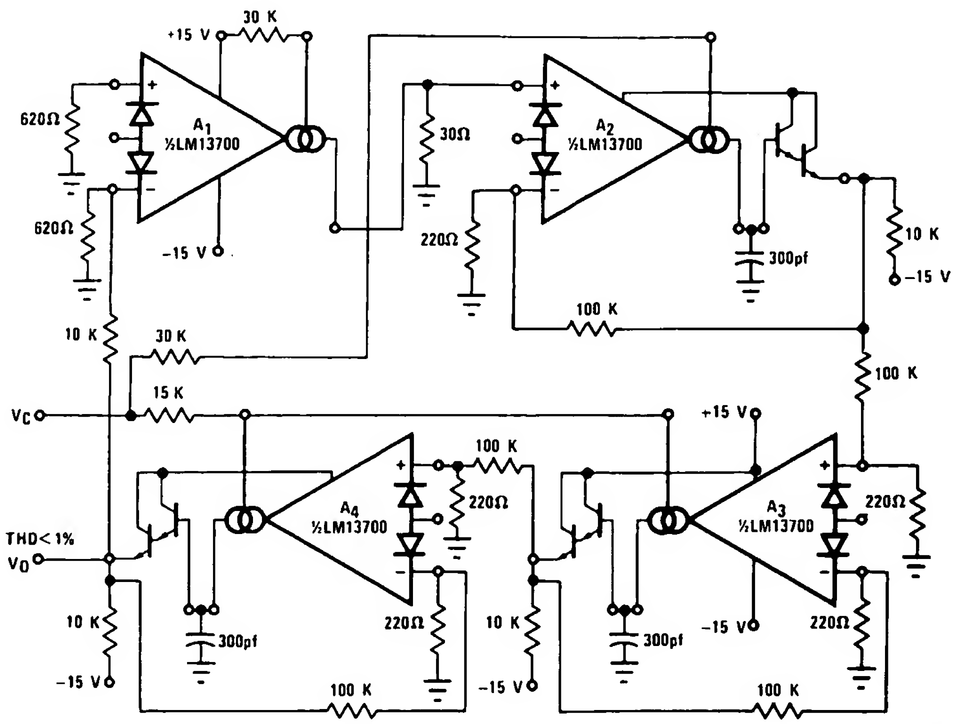


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FIGURE 17. Sinusoidal VCO

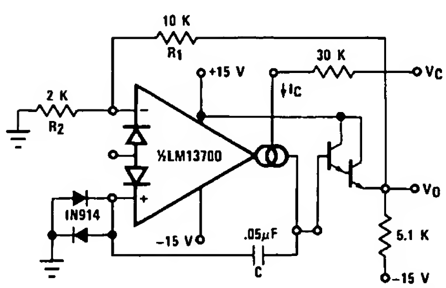

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Figure 18 shows how to build a VCO using one amplifier when the other amplifier is needed for another function.

FIGURE 18. Single Amplifier VCO

Additional Applications

Figure 19 presents an interesting one-shot which draws no power supply current until it is triggered. A positive-going trigger pulse of at least 2V amplitude turns on the amplifier through R_B and pulls the non-inverting input high. The amplifier regenerates and latches its output high until capacitor C charges to the voltage level on the non-inverting input. The output then switches low, turning off the amplifier and discharging the capacitor. The capacitor discharge rate is speeded up by shorting the diode bias pin to the inverting input so that an additional discharge current flows through D_I when the amplifier output switches low. A special feature of this timer is that the other amplifier, when biased from V_O, can perform another function and draw zero stand-by power as well.

Additional Applications (Continued)

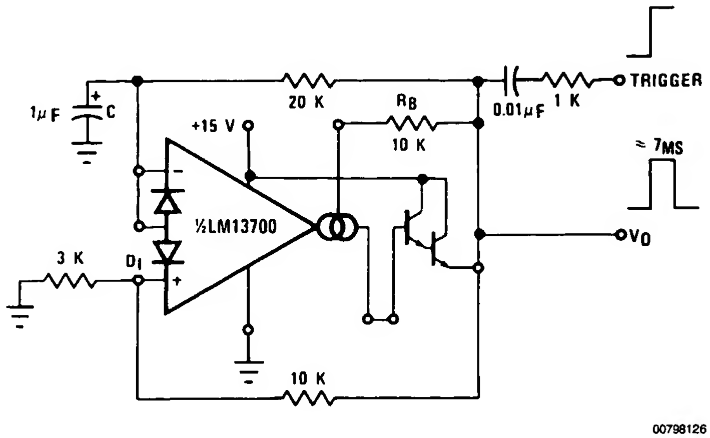


FIGURE 19. Zero Stand-By Power Timer

The operation of the multiplexer of *Figure 20* is very straightforward. When A1 is turned on it holds V_O equal to V_{IN1} and when A2 is supplied with bias current then it controls V_O. C_C and R_C serve to stabilize the unity-gain configuration of amplifiers A1 and A2. The maximum clock rate is limited to about 200 kHz by the LM13700 slew rate into 150 pF when the (V_{IN1}–V_{IN2}) differential is at its maximum allowable value of 5V.

The Phase-Locked Loop of *Figure 21* uses the four-quadrant multiplier of *Figure 6* and the VCO of *Figure 18* to produce a PLL with a ±5% hold-in range and an input sensitivity of about 300 mV.

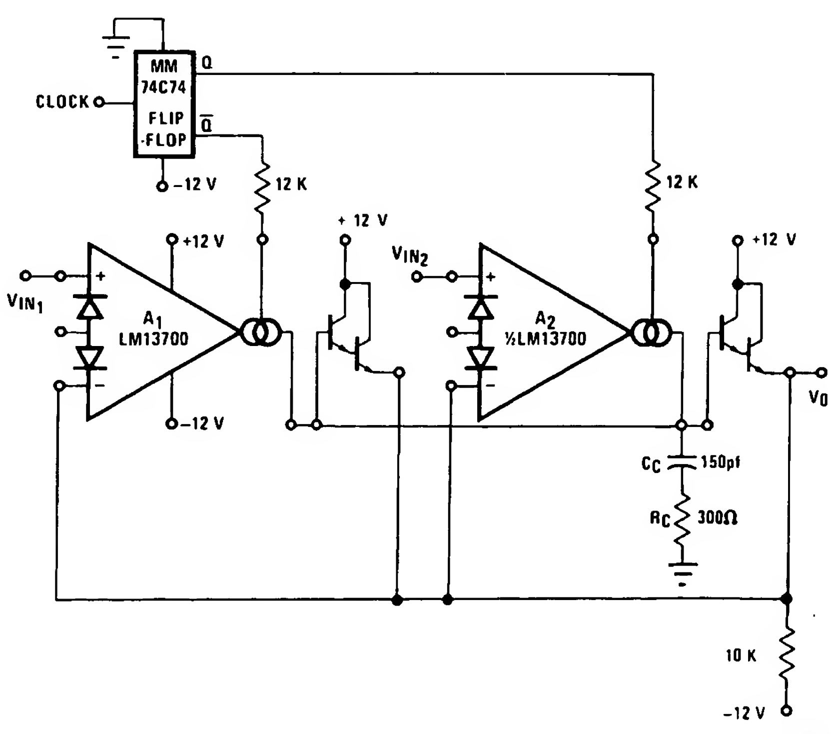


FIGURE 20. Multiplexer

Additional Applications (Continued)

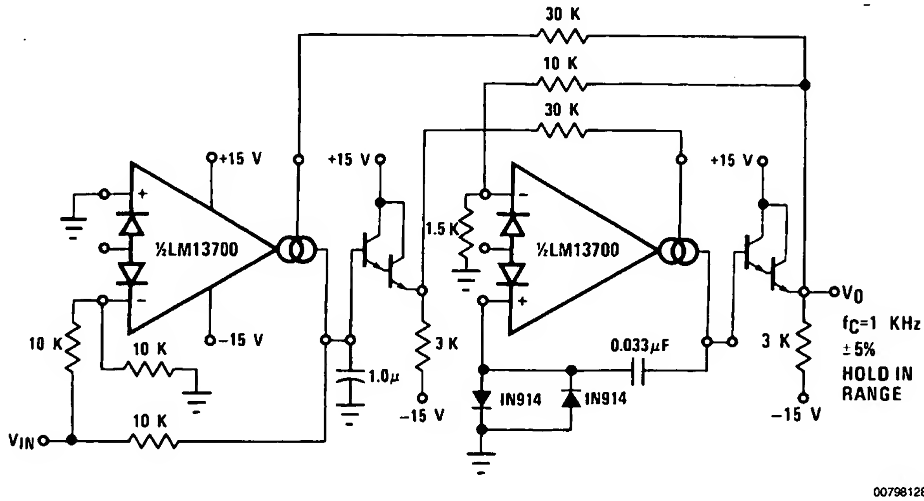


FIGURE 21. Phase Lock Loop

The Schmitt Trigger of *Figure 22* uses the amplifier output current into R to set the hysteresis of the comparator; thus $V_H = 2 \times R \times I_B$. Varying I_B will produce a Schmitt Trigger with variable hysteresis.

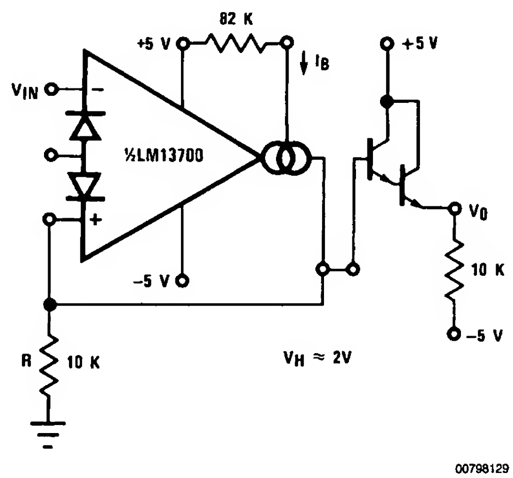


FIGURE 22. Schmitt Trigger

Figure 23 shows a Tachometer or Frequency-to-Voltage converter. Whenever A1 is toggled by a positive-going input, an amount of charge equal to $(V_H-V_L)\ C_t$ is sourced into C_t and R_t. This once per cycle charge is then balanced by the current of V_O/R_t. The maximum F_{IN} is limited by the amount of time required to charge C_t from V_L to V_H with a current of I_B, where V_L and V_H represent the maximum low and maximum high output voltage swing of the LM13700. D1 is added to provide a discharge path for C_t when A1 switches low.

The Peak Detector of *Figure 24* uses A2 to turn on A1 whenever V_{IN} becomes more positive than V_O. A1 then charges storage capacitor C to hold V_O equal to V_{IN} PK. Pulling the output of A2 low through D1 serves to turn off A1 so that V_O remains constant.

Additional Applications (Continued)

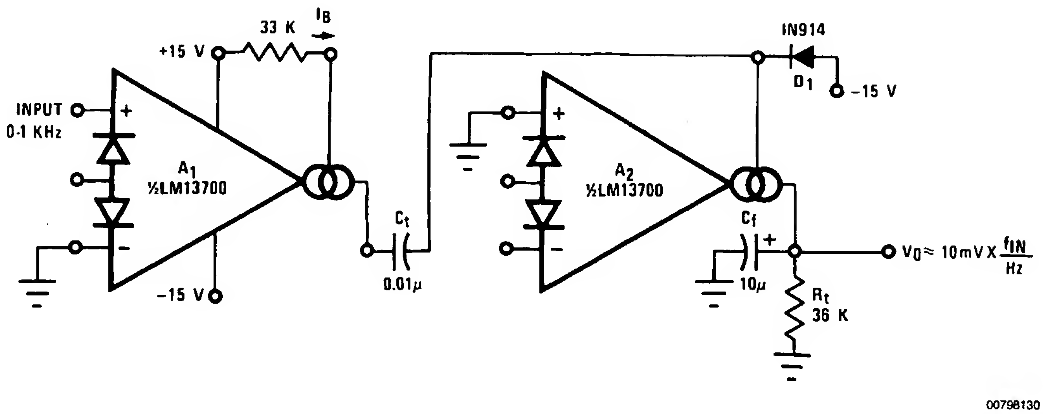


FIGURE 23. Tachometer

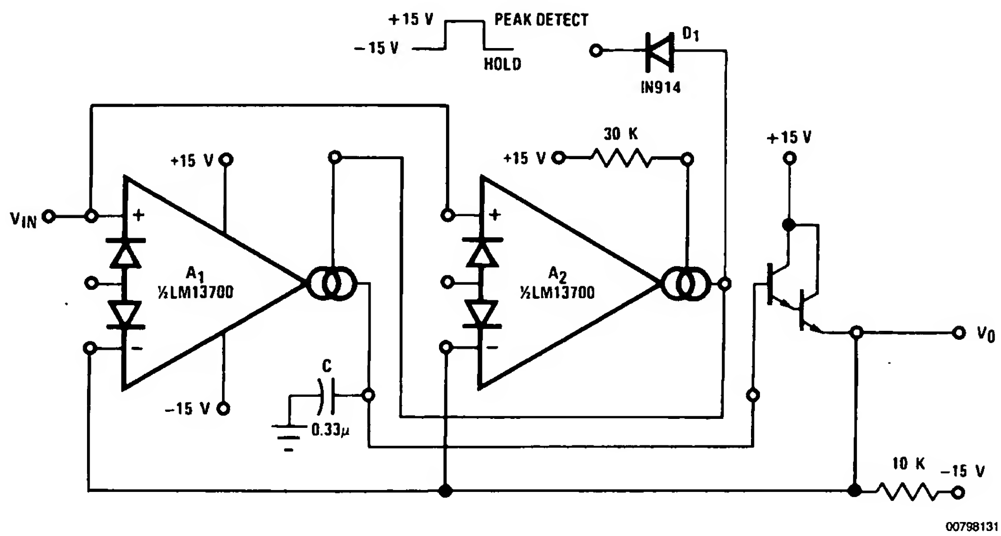


FIGURE 24. Peak Detector and Hold Circuit

The Ramp-and-Hold of *Figure 26* sources I_B into capacitor C whenever the input to A1 is brought high, giving a ramp-rate of about 1V/ms for the component values shown.

The true-RMS converter of *Figure 27* is essentially an automatic gain control amplifier which adjusts its gain such that the AC power at the output of amplifier A1 is constant. The output power of amplifier A1 is monitored by squaring amplifier A2 and the average compared to a reference voltage with amplifier A3. The output of A3 provides bias current to the diodes of A1 to attenuate the input signal. Because the output power of A1 is held constant, the RMS value is constant and the attenuation is directly proportional to the RMS value of the input voltage. The attenuation is also proportional to the diode bias current. Amplifier A4 adjusts the ratio of currents through the diodes to be equal and therefore the voltage at the output of A4 is proportional to the RMS value of the input voltage. The calibration potentiometer is set such that V_O reads directly in RMS volts.

Additional Applications (Continued)

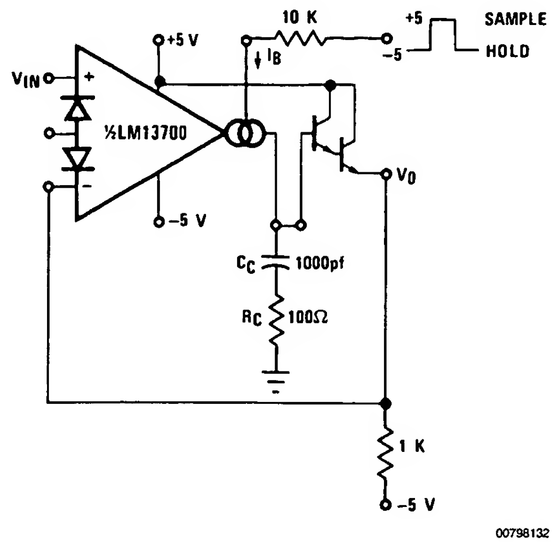


FIGURE 25. Sample-Hold Circuit

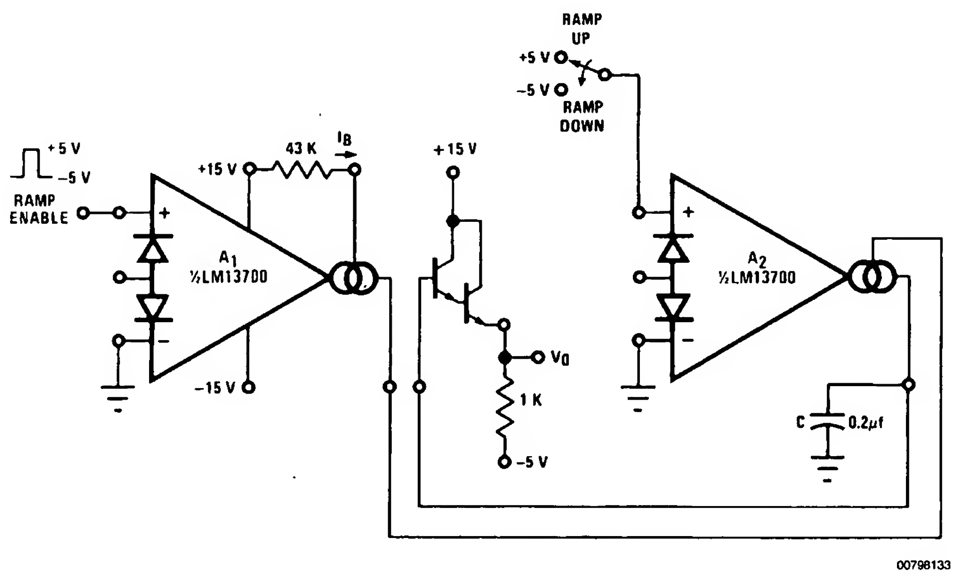


FIGURE 26. Ramp and Hold

Additional Applications (Continued)

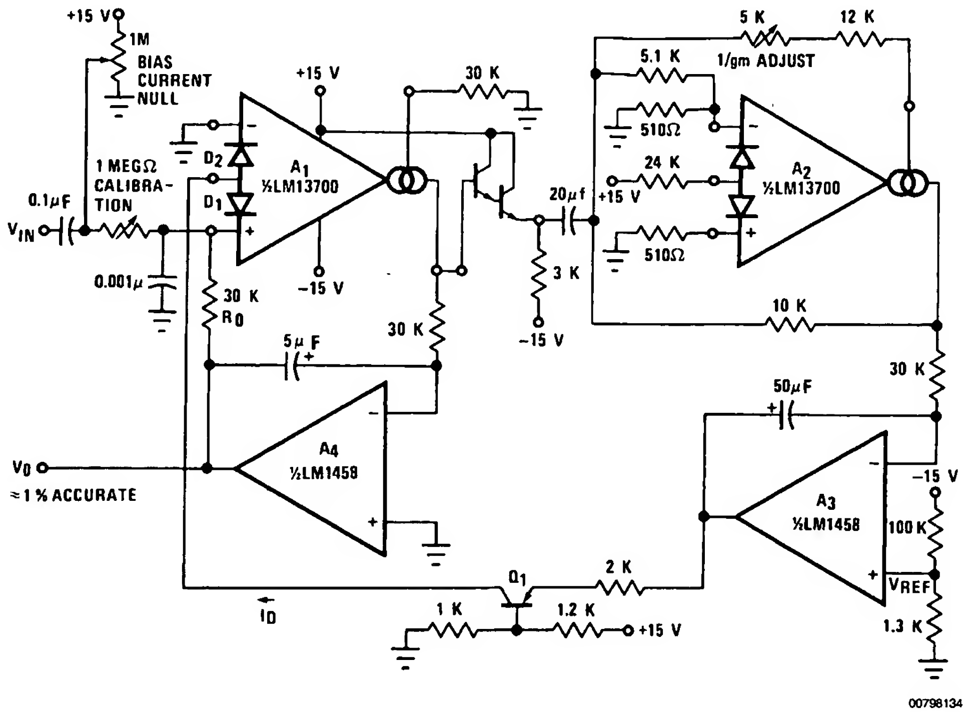


FIGURE 27. True RMS Converter

The circuit of *Figure 28* is a voltage reference of variable Temperature Coefficient. The 100 kΩ potentiometer adjusts the output voltage which has a positive TC above 1.2V, zero TC at about 1.2V, and negative TC below 1.2V. This is accomplished by balancing the TC of the A2 transfer function against the complementary TC of D1.

The wide dynamic range of the LM13700 allows easy control of the output pulse width in the Pulse Width Modulator of *Figure 29.*

For generating I_{ABC} over a range of 4 to 6 decades of current, the system of *Figure 30* provides a logarithmic current out for a linear voltage in.

Since the closed-loop configuration ensures that the input to A2 is held equal to 0V, the output current of A1 is equal to $I_3 = -V_C/R_C$.

The differential voltage between Q1 and Q2 is attenuated by the R1,R2 network so that A1 may be assumed to be operating within its linear range. From *Equation (5),* the input voltage to A1 is:

$$V_{IN}1 = \frac{-2kTI_3}{qI_2} = \frac{-2kTV_C}{qI_2R_C}$$

The voltage on the base of Q1 is then

$$V_B1 = \frac{(R_1 + R_2)\,V_{IN}1}{R_1}$$

The ratio of the Q1 and Q2 collector currents is defined by:

$$V_B1 = \frac{kT}{q}\ln\frac{I_{C2}}{I_{C1}} \approx \frac{kT}{q}\ln\frac{I_{ABC}}{I_1}$$

Combining and solving for I_{ABC} yields:

$$I_{ABC} = I_1 \exp\frac{2(R_1 + R_2)\,V_C}{R_1\,I_2\,R_C}$$

This logarithmic current can be used to bias the circuit of *Figure 4* to provide temperature independent stereo attenuation characteristic.

Additional Applications (Continued)

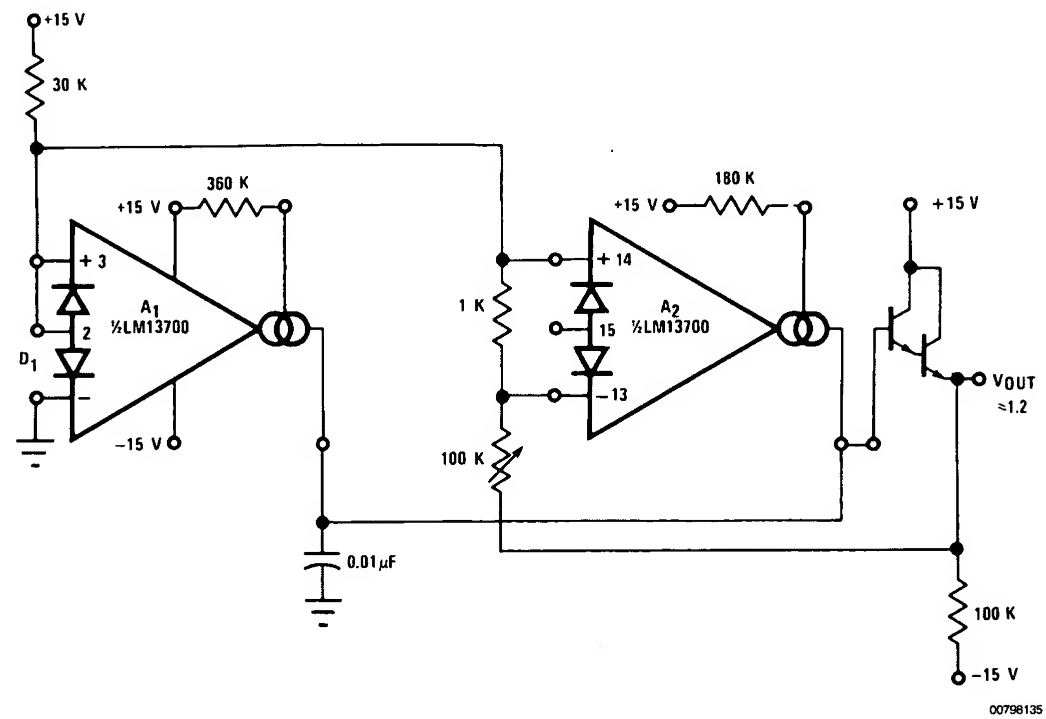


FIGURE 28. Delta VBE Reference

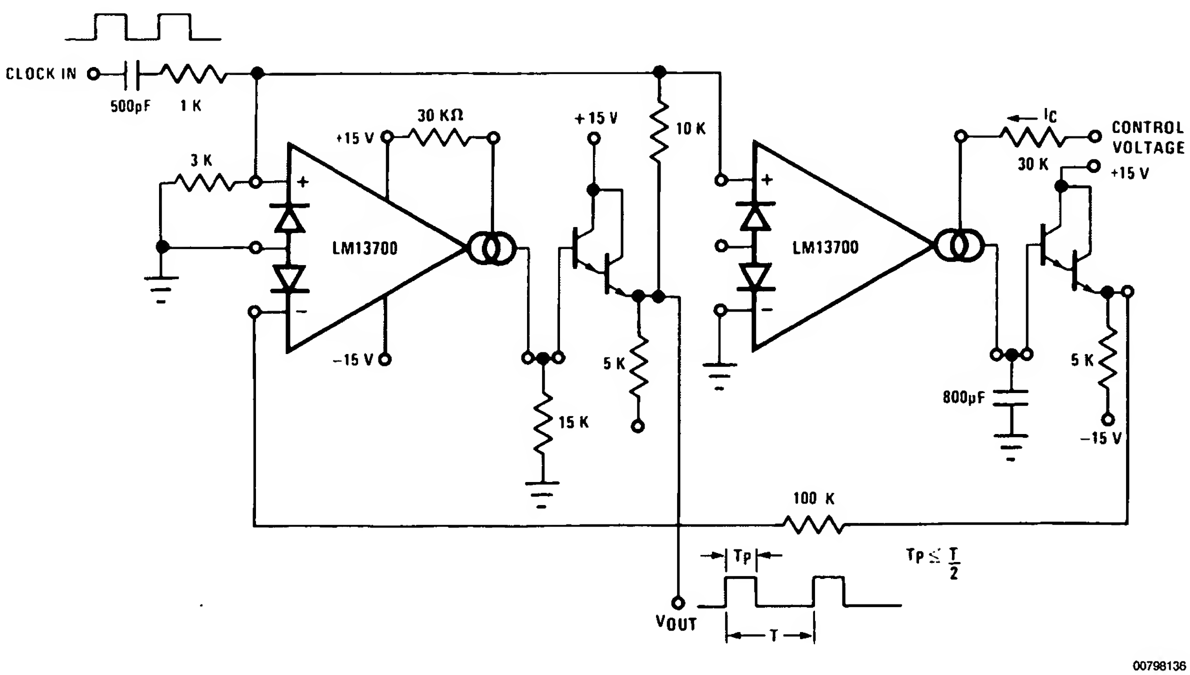
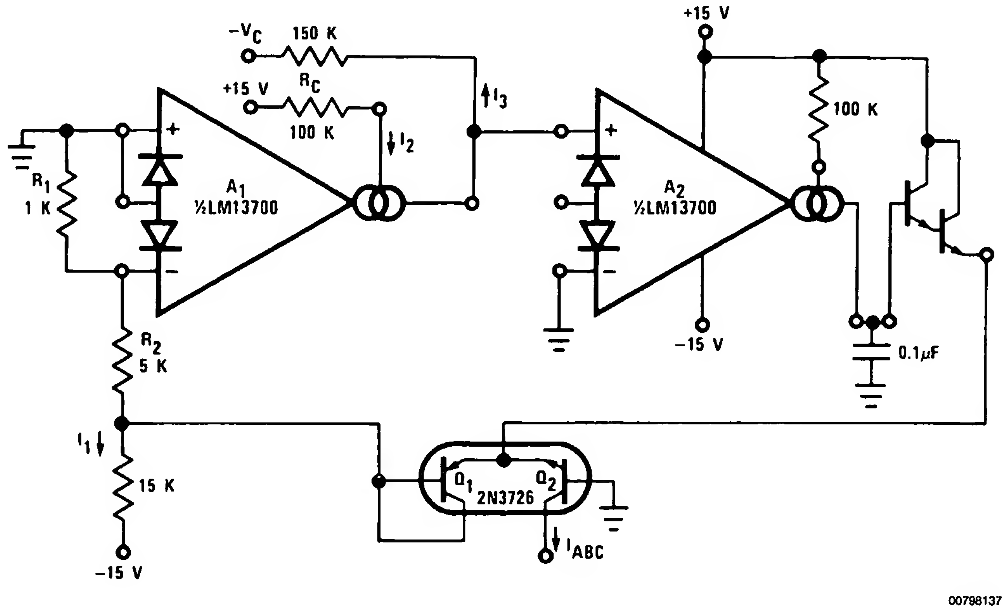


FIGURE 29. Pulse Width Modulator

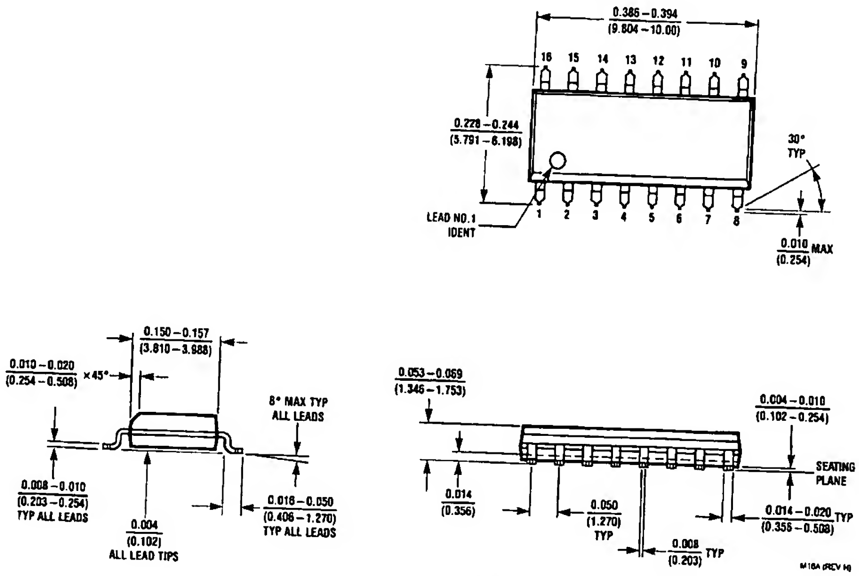
Additional Applications (Continued)



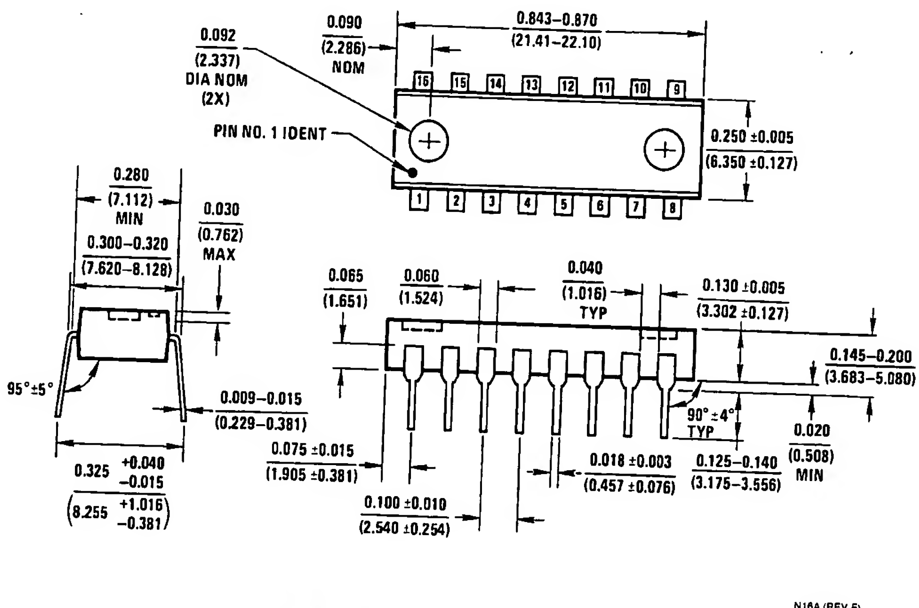
$$I_{ABC} = I_1 \exp \frac{-CI_3}{I_2}$$

FIGURE 30. Logarithmic Current Source

Physical Dimensions inches (millimeters) unless otherwise noted



S.O. Package (M)
Order Number LM13700M or LM13700MX
NS Package Number M16A



Molded Dual-In-Line Package (N)
Order Number LM13700N
NS Package Number N16A

Notes

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